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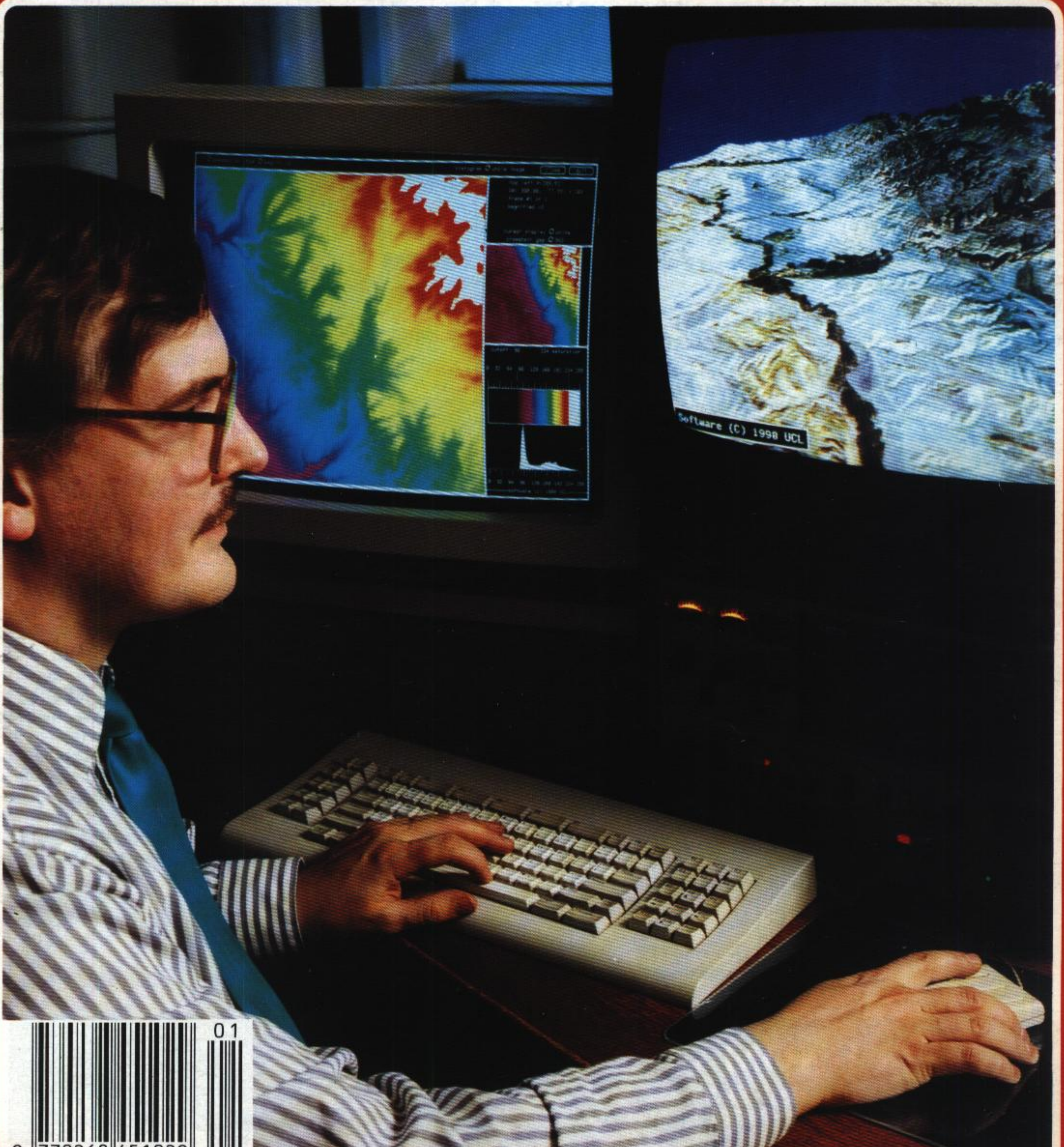
Elektor Electronics

Audio preamplifier (1)
SWR meter
Logic analyser (1)
Multifunction PC I/O card
8031 single-board computer
Error detection & correction



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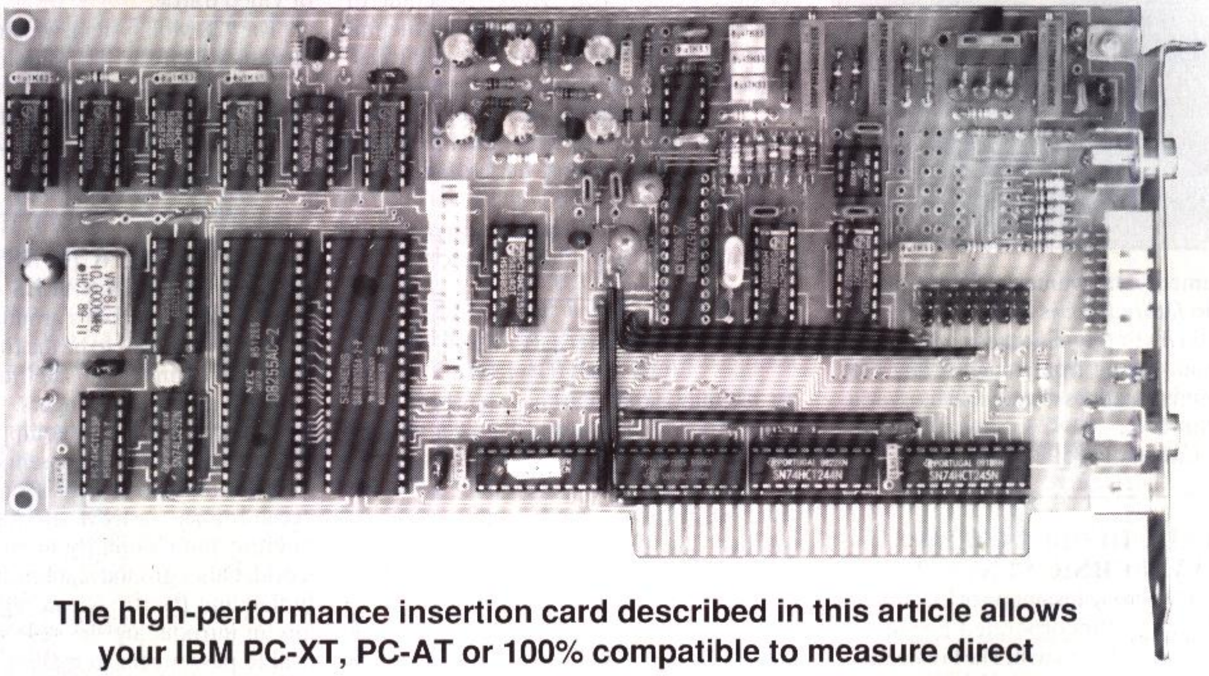
MULTIFUNCTION MEASUREMENT CARD FOR PCs

PART 1: DESIGN BACKGROUND AND CIRCUIT DESCRIPTION



The high-performance insertion card described in this article allows your IBM PC-XT, PC-AT or 100% compatible to measure direct voltage at 12-bit accuracy, as well as frequency and a host of other parameters related to pulse-shaped signals. The accuracy and versatility afforded by the card are of a level associated with much more expensive, industrially rated products. The menu-driven control software developed for this exciting project allows you to keep tabs on up to eight voltages quasi-simultaneously, while up to eight remaining inputs can be used for time-related measurements including frequency, duty factor and pulse duration, not forgetting the event counter. Connected to the sensors and timing devices of your choice, this card turns a PC into a powerful central controller in a complex measurement and control system.

J. Ruffell

ALTHOUGH its ability to deal with numbers is often the main reason for using a PC, it is interesting as well to look at interfacing such a machine to the real world, which, as most of us see it, is analogue rather than digital. Invariably, a PC needs to be fitted with a special interface card to allow it to handle analogue quantities such as voltage, temperature, pressure, or relative humidity. The card described in this article is such an interface. Effectively, it turns your PC into a multifunction measurement station.

The analogue-to-digital conversion function of the card allows analogue quantities in the form of voltages to be translated into their digital equivalents, which are numbers that can be processed and evaluated by the PC. Two inputs are available for this purpose, each fitted with an eight-channel multiplexer for multi-point measurements. The software allows one of these channels to be selected for accurate monitoring and analysis. In addition to this versatile voltmeter function, the card can also function as a frequency meter combined with a pulse/pause (duty factor), a pulse-on/off time meter, and an event counter.

Block diagram

The block diagram in Fig. 1 shows the general structure of the measurement card. A number of buffers and an address decoder are connected between the PC expansion bus on the one hand, and two programmable peripheral interfaces (PPIs) and one analogue-to-digital converter (ADC), on the other. The PPIs take care of the communication between the PC and a number of lines of the ADC (which functions as a voltmeter) as well as the communication with the ICs that form the frequency meter.

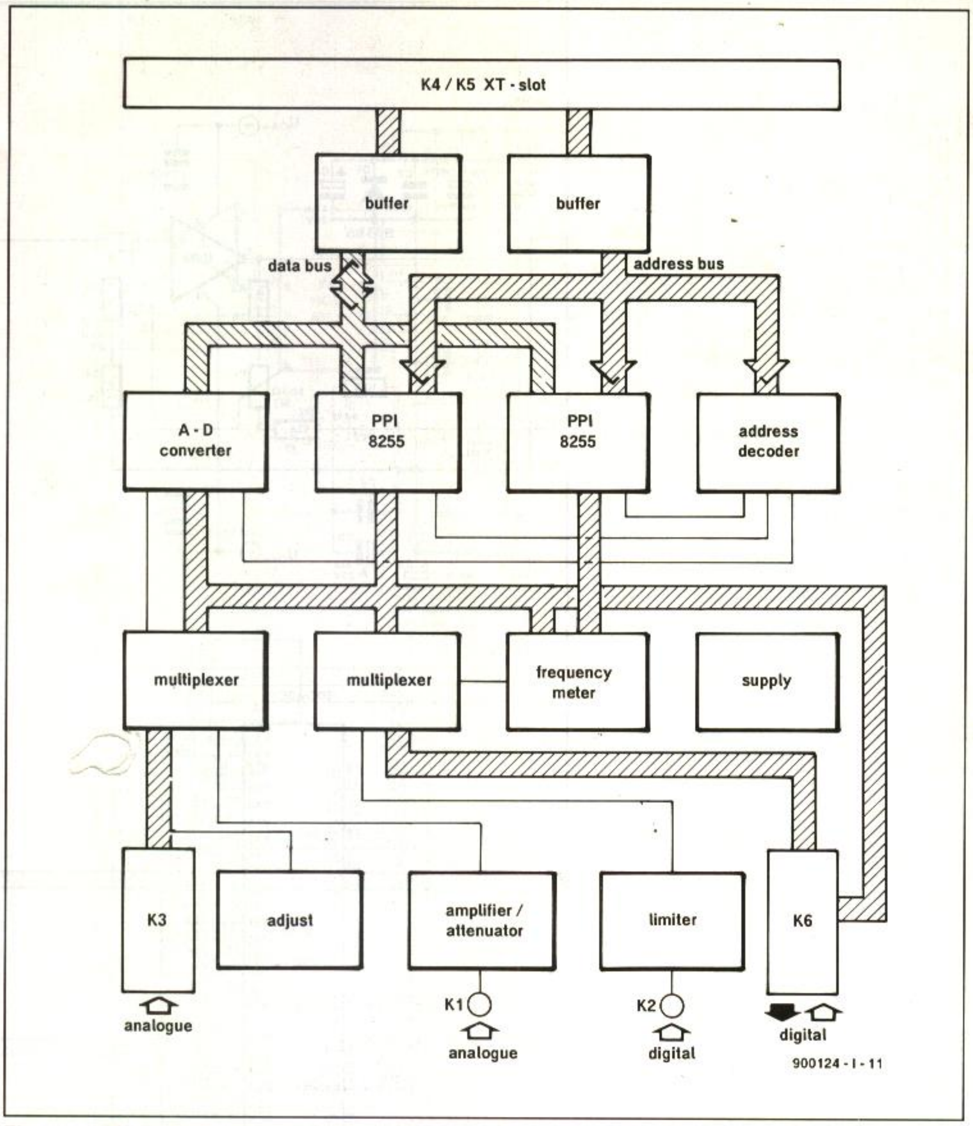


Fig. 1. Block diagram of the PC measurement card, showing the general structure of the design and the main buses used for data and control signals.

MAIN SPECIFICATIONS

- **DC voltmeter**
  - ADC input range: 0 – 5 V
  - ADC resolution: 1.22 mV (12-bits)
  - ADC conversion time: 3 µs
  - Level shifter: 0 – 5 V (input –2.5 V to +2.5 V)
  - 8 multiplexed inputs with 7 optional resistive dividers and 1 adjustable attenuator (0.1 – 300 V, autoranging)
- **Frequency meter**
  - Range: 0.0025 Hz – 10 MHz
  - Max. error: 0.0001%
  - Accuracy: 6 digits
  - Autoranging
  - 8 multiplexed inputs (TTL)
  - Arithmetic period measurement
  - Internal or external reference frequency
- **Event counter**
  - Range: 32 bits
  - Max. count frequency: 10 MHz
  - User-defined trigger edge
- **Pulsetime measurement**
  - Range: 0 – 400 s
  - Resolution: 0.1 µs
  - High/low indication
  - Arithmetic duty cycle measurement

The bus interface merits a few words only, since it has been used in previously published circuits in this magazine. The address and data buffers ensure that the loading of the relevant PC bus lines is kept to a minimum. The buffers are followed by an address decoder formed by a programmable array logic IC (PAL). The PAL outputs are used to give the two PPIs and the ADC their proper addresses in the PC's I/O map. Here, the function of the PPIs is to allow those ICs that are either relatively slow or not really intended for use in a computer environment, to be connected to the PC expansion bus. By virtue of the PPIs, it is possible to apply software control to the multiplexers and the discrete frequency meter, and also to monitor, by means of polling, the status of several components in the system.

The ADC chip used here is suitable for direct connection to a computer bus. Its software selection is accomplished by the $\overline{\text{OUT0}}$ output signal supplied by the address decoder. The EOC (end of conversion) and $\overline{\text{BUSY}}$ signals supplied by the ADC are read via a PPI.

The power supply is symmetrical. The ±12 V supply of the PC is stepped down to ±8 V on the card, and is also used to derive special reference potentials of ±5 V. The digital components on the card are powered direct from the PC's 5-V supply.

Entering into details

Although a block diagram is quite useful to become acquainted with the overall structure of a circuit, a fully detailed circuit diagram is required to understand how the various functions are realized in practice. Hence, the operation of the measurement card will be discussed with reference to the circuit diagram in Fig. 2.

Connectors K4 and K5 (to the left in the circuit diagram) connect the circuit to the XT-slot in the PC. Circuits IC1, IC2 and IC3 provide the necessary buffering. The first, IC1, is a bidirectional type since its task is to buffer the databus, on which two-way communication is required. The other two buffers work in one direction only.

The address decoding is accomplished by IC7, a PAL. This IC, which is supplied ready-programmed, has been used before in a PC extension card project, see Ref. 1. Apart from the three selection signals, the PAL also supplies the reset signal for the ADC interrupt bistable, IC6A, and the enable signal for bus buffer IC1. The base address of the card can be set with the aid of jumper JP1 (position A: 300H; position B: 310H). Table 1 shows the functions of the addresses assigned to the card.

The analogue circuitry is found back at the top of the circuit diagram. To the right we see an analogue multiplexer, IC12, which is used to control a stepped attenuator built from discrete components (R23 – R32). The attenuator is followed by a chopper-stabilized amplifier (Ref. 2), IC11, whose output signal is fed to a second analogue multiplexer.

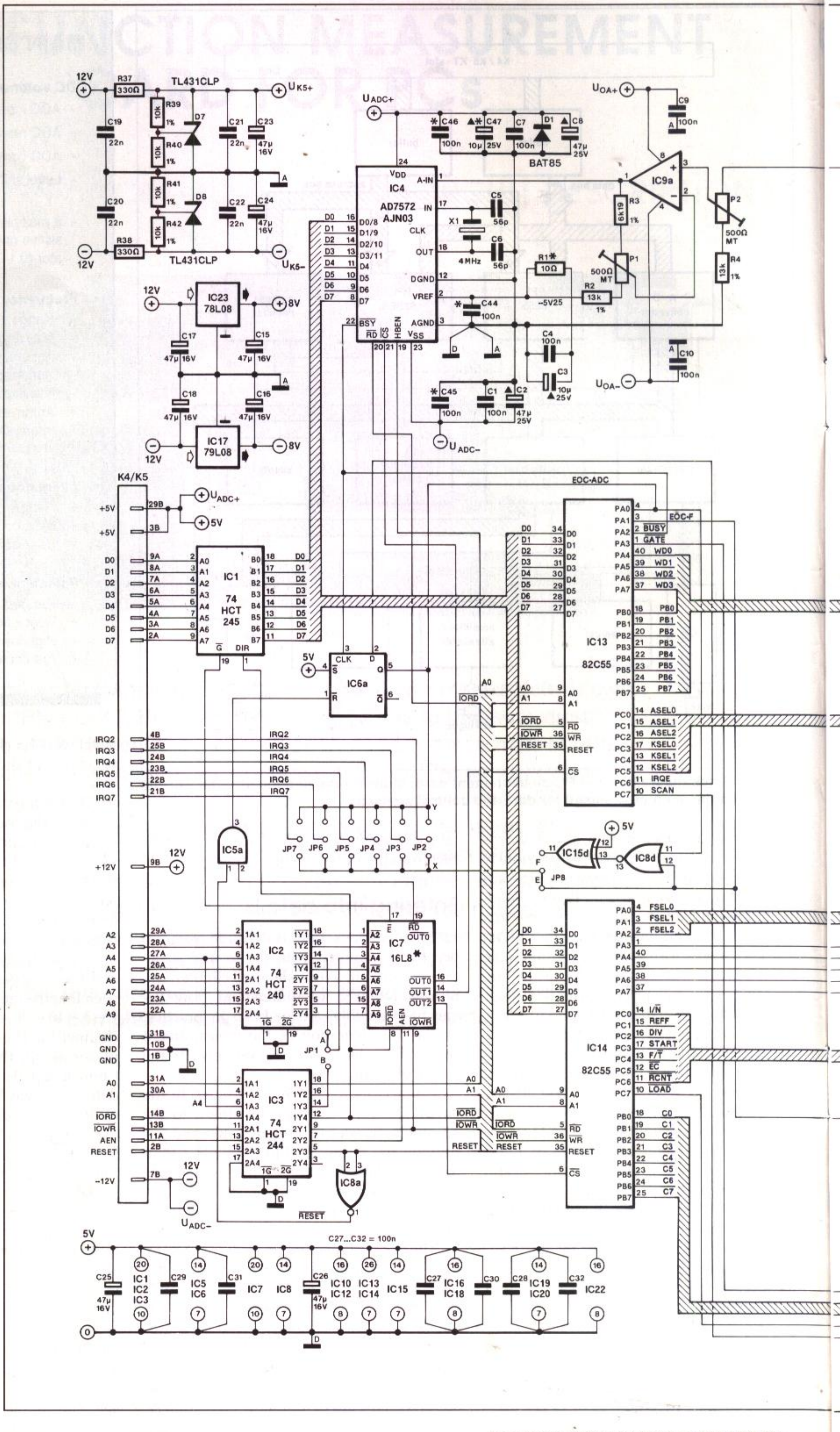
The component values in the attenuator and the chopper opamp circuit are chosen such that the measurement ranges 0.1 – 0.3 – 1 – 3 – 10 – 30 – 100 – 300 V are created on channel I0. The other inputs of the multiplexer accept the signals on connector K3. Control signals ASEL0, ASEL1 and ASEL2 determine which of the eight signals at the multiplexer inputs is fed to ADC IC4, via buffer IC9B and level shifter IC9A. The ASEL signals are supplied by one of the PPIs, IC4. The arrangement allows a maximum of eight signals to be applied sequentially to the ADC. In the current circuit configuration, input I1 (pin 1 of connector K3) is used to calibrate the ADC. The calibration circuit can be disabled by fitting jumper JP10, when input I1 may be used as an additional measurement input.

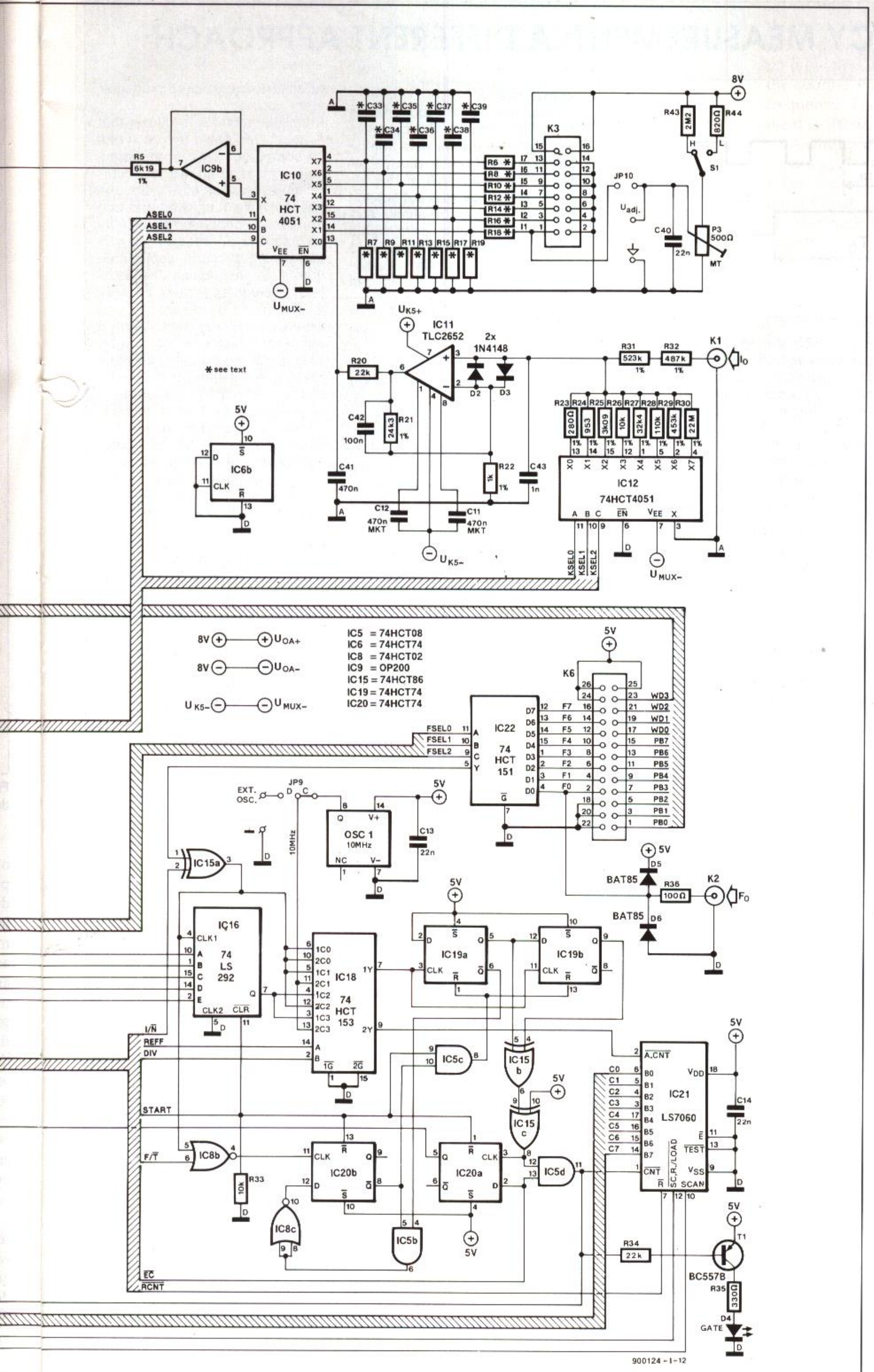
It should be noted that inputs I1 to I7 are **not** fitted with a protective network, so that any attenuator used on these inputs must include a form of protection against overvoltage. The simplest way to accomplish this is probably to fit a pair of anti-parallel diodes. Note, however, that these diodes are not required, even undesirable (because of their leakage current), when measuring sensor voltages that can not exceed the operating area of the multiplexer (±5 V). Depending on the desired measurement range, you must calculate the values of R6 to R19 yourself. It will be clear that high-value resistors are preferred to keep ground currents in check. Each resistor junction in the attenuator may be decoupled with a capacitor (C33 – C39).

Level shifter IC9A allows negative voltages to be measured. With the aid of the buffer/amplifier ahead of it, the level shifter enables voltages between −2.5 V and +2.5 V to be measured at an accuracy of 1.22 mV.

The heart of the circuit is a Type AD7572A analogue-to-digital converter from Analog Devices. This IC is supplied in four versions. The -A suffix version used here is the fastest with a conversion time of only 3 µs. It requires a 4-MHz quartz crystal, X1, and a wire link in place of resistor R1. If you feel you can make do with a longer conversion time, you may want to use one of the slower, less expensive, versions of the AD7572. These, however, require a lower quartz crystal frequency, say, 1.25 MHz for a version with a conversion time of 10 µs. At the lower clock frequencies, it may be necessary to adapt the values of capacitors C5 and C6 to ensure that the oscillator starts reliably. When the older version of the AD7572A, the AD7572, is used, resistor R1 must be fitted at its appropriate position on the PCB.

The ADC used has an internal reference that supplies a voltage of –5.25 V at pin 2. This voltage is used for the on-chip A-D conversion circuitry as well as for the pre-





viously mentioned level shifter. The voltage level at pin 22 of the ADC indicates whether or not the A-to-D conversion process is complete. This allows the computer to monitor the conversion status with the aid of the PPI, although this type of check is really only useful when using either a slow ADC or a very fast computer. In most cases, a conversion time of 3-μs is too fast in any case for the computer to execute the relevant start instructions, and evaluate the results. This is particularly so when the control program is written in a higher language.

The above discussion may suggest that a fast ADC is extravagant and a superfluous luxury. Bear in mind, however, that a fast ADC can take more samples of a measurand within a short period, allowing the software to calculate the correct value on the basis of the sampled values. This type of software-based filtering effectively eliminates the adverse effects of spurious pulses and noise on the measurement results, and brings true 12-bit resolution within easy reach.

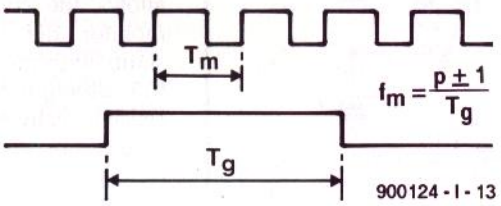
A different principle

A fairly unusual principle is adopted for the measurement of frequency. In most frequency meters, the frequency of the input signal is measured by counting the number of leading or trailing pulse edges that occur within a fixed period, called the gate time. In this system, the measurement error increases considerably as the number of counted pulses decreases within

Fig. 2. Circuit diagram of the multifunction measurement card for PCs. The main functions of the card are a voltmeter and a frequency/pulse meter. Both are implemented with a very high accuracy of 12 bits and 1 p.p.m. respectively.

FREQUENCY MEASUREMENT: A DIFFERENT APPROACH

1



In all probability, the simplest way of measuring frequency is counting the number of periods (p) of a signal within a fixed gate time (T_g). Unfortunately, this method has one important disadvantage: the so-called ±1-error, which means that the last pulse in a measurement or conversion is always undetermined. Owing to this error, the accuracy of the simple measurement drops inevitably with the input frequency (see Fig. 1). Although the effect of the ±1-error can be reduced by increasing the gate time, measurement times rapidly rise to excessive lengths when dealing with low frequencies. For example, a 6-digit (0.0001%) accurate measurement of a 1 Hz signal would require a gate time of 277.8 hours! Clearly, this is a totally impractical measurement time.

In addition to lengthening the gate time, the accuracy may also be improved by first multiplying the frequency to be measured with a known reference frequency, and subsequently measuring the product frequency. The result may then be used to calculate the original input frequency. Unfortunately, multiplying signals whose frequencies differ by a considerable amount is not so simple, so that this measurement method is far from ideal.

A much more efficient method to determine the frequency of a signal is based on period time measurements, where a circuit counts pulses supplied by a stable reference during one period of the measurand (see Fig. 2). In this way, the measurand switches the reference on and off. As in the 'simple' measurement system, the accuracy achieved is dependent on the frequency to be measured — since a fixed reference frequency is used, the effect of the ±1-error rises with the frequency to be measured. This is caused by the 'gate time' becoming shorter with higher input frequencies, so that fewer periods of the reference source can be counted. Remember, the effect of a ±1-pulse error is ten times greater with 100 counted pulses than with 1,000 counted pulses.

The accuracy of this type of measurement may be increased by lowering the input frequency. This is fairly simple to accomplish by dividing the input signal by a known factor. As illustrated in Fig. 3, the real frequency may then be computed on the basis of the lowered frequency and the divisor.

Briefly recapitulating, the method of counting reference periods has two advantages over counting input signal periods:

- the time required for the measurement equals the period time of the (scaled-down) input frequency, and thus remains within acceptable values;
- the fact that signal division may be applied instead of signal multiplication allows a simpler circuit to be used.

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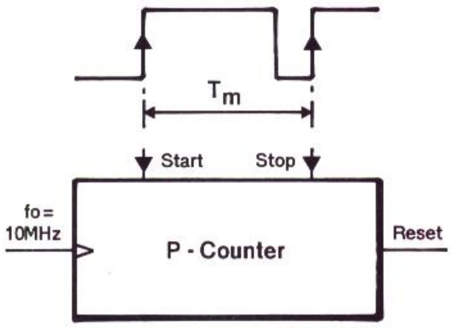

In practice

Among the target requirements of the PC measurement card was a maximum permissible measurement error of 1 p.p.m. (or 1×10^{-4}%) as a result of the ±1-error. This value was not set just like that — it can be shown that it is the maximum permissible error if the measurement is to have 6-digit accuracy. Mathematically, this requirement can be expressed as:

$$f_m \leq 1\times10^{-6} f_o \qquad [1]$$

where f_m is the frequency to be measured, and f_o the frequency supplied by the reference source. Since the reference operates at 10 MHz in this case, equation [1] rules a maximum value of f_m of 10 Hz. This means that the scaled-down frequency, f'_m, must comply with equation [1]. In other words: $f'_m < 10$ Hz.

In terms of hardware, the input signal is scaled down by IC16, a 74LS292. This IC allows a divisor 2^n to be programmed, where n may take discrete values from 2 to 31. From this, it can be shown that n must comply with

$$n > (\log_{10} f_m / 10) / (\log_{10} 2) \qquad [2]$$

to ensure the required measurement accuracy.

From equation [2] it follows that f_m must be known before n can be computed, while at the same time n must be known for accurate measurement of f_m. Although this looks like a chicken-and-egg problem, acceptable results may be obtained for the setting of n by performing a rough approximation of f_m. In practice, the problem is solved as follows. First, f_m is estimated with the aid of a preliminary measurement, carried out with a factor, n, of nought. This factor results in the shortest measurement time. If the result of the preliminary measurement is greater than 10 Hz, the frequency is entered as fm in equation [2]. Next, the computer calculates the optimum value of n, and sets the programmable divider to the required divisor.

Next, the real measurement follows. The result of it (f'_m), multiplied by 2^n provides the frequency of the input signal, at the required accuracy of 6 digits. When the preliminary measurement results in a value of f_m smaller than 10 Hz, the 'first go' is already sufficiently accurate, and thus obviates a second measurement. ❑

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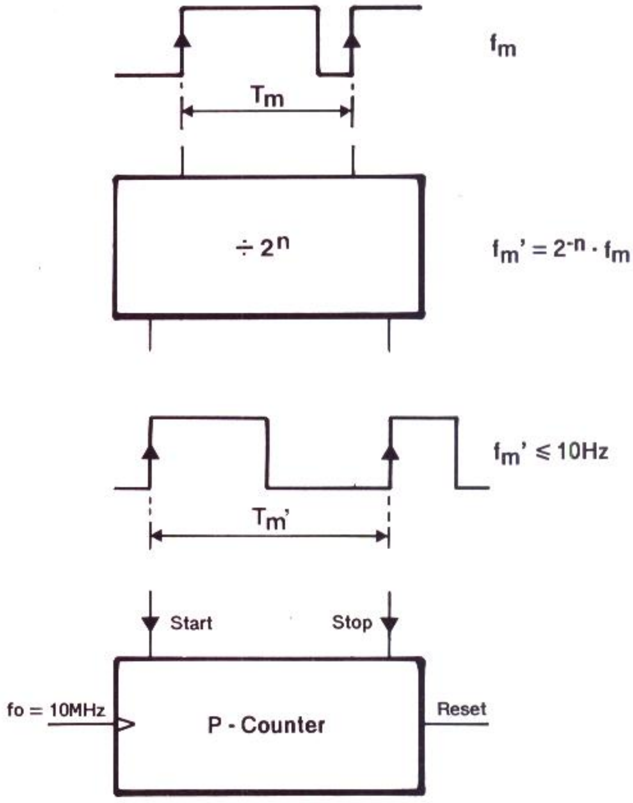

Address	Function
3x0	Read ADC lower byte
3x1	Read ADC upper byte
3x2	Read ADC lower byte
3x3	Read ADC upper byte
3x4	Port A IC13 (input)
3x5	Port B IC13 (not used)
3x6	Port C IC13 (output)
3x7	Control IC13
3x8	Port A IC14 (output)
3x9	Port B IC14 (input)
3xA	Port C IC14 (output)
3xB	Control IC14
JP1 = A:	x = 0
JP1 = B:	x = 1

All addresses in hexadecimal

Table 1. Functions of addresses assigned to the measurement card.

the gate time. This principle is, therefore, not practical for the measurement of very low frequencies, since these would require gate times of minutes, or even hours, to ensure the necessary accuracy. Hence, the present PC measurement card is based on a totally different principle for measuring frequency. This principle, which is explained separately on the previous page, does not have the above disadvantage of low accuracy at low frequencies.

As with the voltage meter circuit, a multiplexer (IC22) is fitted between the inputs and the actual measurement circuit. One of these eight inputs is provided with a protection circuit, and taken to a pin on connector K2. The remaining seven are connected to K6 only. From the output of the multiplexer, the selected signal is fed to XOR gate IC15A. This gate may be used to invert the signal under the control of the $I/\overline{N}$ line. The output of the XOR gate is connected to a 2^n divider, IC16, a multiplexer, IC18, and a NOR gate, IC8B. The multiplexer thus has three input signals: the input signal (in true or inverted form), the input signal divided by 2^n, and the 10-MHz clock signal supplied by oscillator block OSC1. Control signals REFF and DIV on the A and B inputs determine which input signals appear at the multiplexer outputs, 1Y and 2Y. The multiplex configuration is shown in Table 2.

State	DIV	REFF	1Y	2Y
Event count	n.a.	L	n.a.	f_m
Pulsetime	L	H	f_m	f_o
Frequency:				
first go	L	H	f_m	f_o
real	H	H	f'_m	f_o

f_m = frequency of input signal
$f'_m = f_m/2^n$
f_o = 10-MHz reference frequency
n.a. = not applicable

Table 2. Control signal programming to select the various functions of the card.

Circuits IC19A and IC19B form a two-bit shift register that is used to detect exactly one period during frequency measurements. Following a reset pulse, both Q outputs are low. They are also low after two leading edges of the clock signal. As long as only one output is low (during one period of the clock signal), the clock signal is passed to a 32-bit counter, IC21 (a Type LS7060), via XOR gates IC15B-IC15C and AND gate IC15D. Note that the counter is an LSI chip, not a device from the 74LS series. Simultaneously with the enabling of the 32-bit counter, the gate LED, D4, lights to indicate that a measurement is being performed.

The measurement is started in a relatively simple manner. First, the 32-bit counter is cleared with the aid of the $\overline{\text{RCNT}}$ signal. The START signal goes low, and resets the other registers. Next, START goes high, so that the shift register enables the counter during one period.

After one period of the input signal, the signal EOC-F is actuated (START = high) via bistable IC20A. Next, the counter can be read via IC14 with the aid of signals SCAN and $\overline{\text{LOAD}}$.

A relatively simple arithmetic operation allows the period to be deduced from the results of the frequency measurement. In addition, the pulse on-time can be measured. Since it possible to invert the input signal, it is a relatively simple matter to measure both the 'high' time and the 'low' time of the input signal. The circuit based on bistable IC20B and its associated gates perform this task elegantly. The PC uses the $F/\overline{T}$ signal to select between frequency measurement and pulse time measurement, while the logic level on the $I/\overline{N}$ line determines whether the measured time corresponds to the 'low' or the 'high' part of the input pulse. When $I/\overline{N}$ is logic low, IC15A will not invert the input signal. Consequently, IC19A is reset by the first leading edge after the START command. From that moment on, counter IC21 remains enabled until IC19A is reset by the next pulse transition, which is trailing. This resetting takes place via IC20B.

During the measurement, a NOR gate, IC8B, propagates the measurand to the clock input of bistable IC20B, which is thus clocked at each trailing edge of the input signal. By virtue of IC5B, this clock signal is only effective with IC19A set, when a logic one is applied to the input of IC20B. Once IC20B is set, IC19A is reset, and the counter disabled. Because the $\overline{Q}$ output is fed back to the data input of the bistable (via IC5B), the state of IC19A and IC21 is frozen until the next start pulse occurs. The appearance of the EOC-F signal at the Q output of IC20A indicates that the measurement is complete. Next, the program reads the counter state. The pulse 'low' or 'high' time is simple to compute since it is the counter value divided by the reference frequency. The resolution of this measurement is 100 ns.

As indicated above, the 'high' time of the input signal is measured with $I/\overline{N}$ not active. When this signal is active, IC5A inverts the input signal, and the 'low' time is measured like the 'high' time, as explained in the previous paragraphs.

For the pulse-related measurements a 10-MHz reference frequency is used. This frequency is supplied by a quartz crystal block, OSC1. Jumper JP9 allows an optional, external, oscillator to be connected.

A feature of the card that has not been discussed so far is its ability to function as an event counter. In this mode, the input signal is connected to the clock input, pin 2, of the counter, via IC18. The counter will count pulses as long as $\overline{\text{EC}}$ (event count) is active, i.e., logic low. This is arranged by the software. In contrast to the situation with the other measurement modes, the event counter mode requires the START signal to remain inactive. To complete the story, we must mention that the signal $I/\overline{N}$ determines which pulse edge of the input signal is used to advance the counter.

In the centre of the circuit diagram we find six jumpers labelled JP2 to JP7. These jumpers allow the interrupt signal supplied by the card to be connected to one of the interrupt lines on the PC expansion bus. As usual with PC insertion cards, due care must be taken to use a free as well as appropriate interrupt line. In most IBM PCs, the interrupt line assignment is as follows:

IRQ2:	reserved
IRQ3:	COM2: (serial I/O)
IRQ4:	COM1: (serial I/O)
IRQ5:	hard disk
IRQ6:	floppy disk
IRQ7:	LPT: (parallel I/O)

The measurement card may be set to use IRQ3 without problems if the PC is a type with only one serial port (COM1 on IRQ4). Where IRQ3 is in use by COM2:, it is nearly always possible to move the measurement card to IRQ2.

The IRQ line assigned to the measurement card must be set in hardware as well as in software. The software setting is accomplished by modifying one line in the configuration file, ADCF.CFG. Further details on the use of interrupt lines are provided in the READ.ME file on the diskette.

Finally, jumper JP8 allows us to either combine the interrupts from the frequency meter and the ADC (jumper position F), or use them separately (jumper position E). ■

Next month's second and final instalment of this article will deal with the construction and practical use of the measurement card, and with the control software developed for it.

MEASUREMENT TECHNIQUES (3)

by F.P. Zantis

Following the discussion in our previous issue on the measurement of voltage, this month's instalment deals with the measurement of current and power.

STRICTLY speaking, each and every moving-coil or moving-iron measuring instrument is a ammeter. The deflection of the pointer is proportional to the level of the current flowing through the instrument. However, only a small number of them have been constructed specially for the measurement of current. In most electronic laboratoria and workshops, multimeters are habitually used for measuring current.

The full-scale deflection of the instruments used in multimeters varies roughly from 50 µA to 500 µA. Since the levels of the currents to be, or being, measured are normally much higher than those values, the metering range must be extended. This is usually done by shunting the meter with an appropriate resistance through which the larger part of the current flows. When a metering range is selected in a multimeter, a different shunt resistor is switched in parallel with the meter—see Fig. 15.

Current measurements by digital meters are transformed to voltage measurements, for which shunt resistors are also required. The potential drop measured across the appropriate resistor is proportional to the current through it. Here again, the ammetering range is extended by the use of a number of resistors.

The measurement of alternating currents is subject to errors, tolerances and problems that are similar to those experienced in measuring alternating voltages. For instance, the measurement is true for only one frequency. Also, the true r.m.s. value is indicated only for a truly sinusoidal current. However, there are, fairly expensive, so-called true-r.m.s. multimeters that indicate the r.m.s. value irrespective of the waveform of the measurand.

Errors in current measuring

Errors in current measurements occur because of the frequently not very high accuracy of the instrument or through the effect the instrument has on the measurand. As in voltage measurements, the internal resistance of the instrument causes a measuring error. To measure a current, the circuit through which it flows has to be opened to enable the meter to be inserted in series. That does, of course, increase the total resistance in the circuit by an amount equal to the internal resistance of the measuring instrument. The higher the internal resistance, the greater the measuring error. For instance, if a circuit has a resistance of 3 Ω and a potential of 12 V is applied across it, a current of I = U / R = 12/3 = 4 A flows. If now a meter with an internal resistance of 0.5 Ω is inserted into the circuit, the current will be 12/(3+0.5) = 3.43 A. That is a measuring error of no less than 17%! The higher the sensitivity of the measuring instrument, the smaller its internal resistance and, consequently, the error.

Indirect current measurement

The level of a current is determined more accurately when the voltage drop it causes across a resistance is measured. The error that occurred in the example in the previous paragraph then becomes negligibly small. The level of the current through the resistance is calculated with the aid of Ohm's law. This method of indirectly measuring the current is normally far preferable over the direct method. Furthermore, it has the advantage that the circuit under test need not be broken into, which, especially in the case of a printed circuit, may be fairly difficult if not impossible.

On the other hand, a direct measurement of current with a simple multimeter may solve a seemingly insoluble difficulty in measuring a voltage. Consider, for instance, the circuit in Fig. 16 where, to enable the operating point to be set, the drop across the collecter resistor must be measured. To do this with a 20 kΩ/V multimeter, a fairly large measuring error would be caused by the internal resistance of the instrument. Here, it is far better to measure the current through the resistor as shown, always provided that the internal resistance of the meter is sufficiently small. The voltage across the resistor is, by Ohm's law, the product of current and resistance.

Note that in the previous paragraph it is not possible to measure the base current, since that is too small to be measured with the usually available instruments.

To estimate the measuring error as precisely as possible, it is necessary to know the internal resistance of the ammeter at the selected metering range. Unfortunately, manufacturers only specify this for the voltage ranges. It is, therefore useful to know how to determine it for the current ranges. This may be done by measuring the current from a constant-current source. Then, successively connect a number of resistors of different values in parallel with the meter inputs as shown in Fig. 17. When the meter indicates half the level of current measured before any resistance was connected in par-

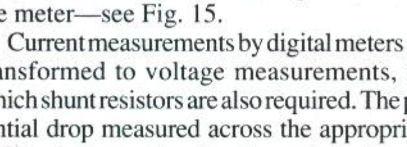

Fig. 15. The metering ranges of current meters are extended by placing resistors called shunts in parallel with the meter.

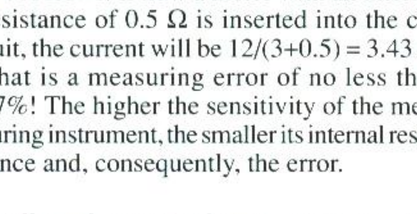

Fig. 16. In difficult cases, the direct method of current measurement may be used to indirectly measure vol-ages to reduce the measurement error.

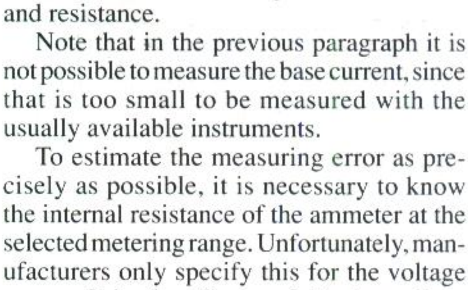

Fig. 17. Multimeter with external shunt resistor.

allel, the internal resistance of the meter is equal to the value of the resistor then shunting the meter inputs.

Extending the metering range

Some multimeters have, in addition to several fused current ranges of up to 2A, an unprotected range of up to 10 A, or even 20 A. In principle, however, most multimeters may be used for measuring high currents, but this entails extending their metering range(s).

The current range(s) of a multimeter may be extended by the same method as used above to determine the internal resistance. If, for instance, a resistor whose value is equal to the internal resistance of the instrument is connected across the input terminals, half of the measurand flows through it. In other words, the indicated value must be multiplied by 2 to obtain the true value of the current. When two such resistors are connected across the input terminals, two thirds of the measurand flows through them and the meter reading must be multiplied by 3. With the well-known formula $P = I^2 R$, the power dissipated in the shunt resistor(s) may be calculated. Unfortunately, resistors with the required power rating are normally available only in the 10% tolerance range, which does not bode well for accurate measurements. A further drawback is that the required resistance values are normally not available as standard resistors, so that series-parallel combinations must be used. An aid to calculating the needed resistance values will be given in the form of a small basic program later in the series

Current measurement with an oscilloscope

For measuring large currents, the indirect method is normally much easier to realize. When the circuit under test has no current-carrying resistor across which the voltage that determines the current can be measured, one must be added. For that purpose, any resistor may be used whose value is small compared with that of the overall circuit. It is inserted into the loop whose current is to be measured and the voltage across it measured. The current is then easily calculated with the aid of Ohm's law. For certain values of resistance, the calculation is very simple. If, for instance, a 1-Ω resistor is used, the value of the measured voltage in volts is equal to the current in amperes. Also, if a 0.1-Ω resistor is used, the indicated value in volts must be multiplied by 10 to give the current in amperes. With this method it is possible to use almost any oscilloscope as an ammeter; even current peaks can be detected—see Fig. 18.

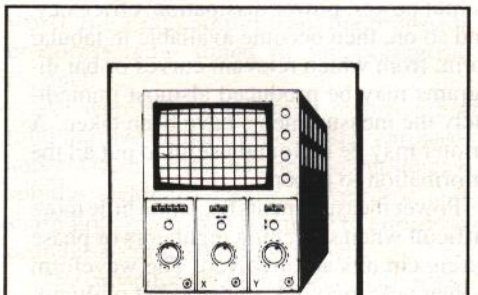

Fig. 18. An oscilloscope may be used for indirect measuring of current.

Even in indirect measuring of current, the tolerance of the shunt increases the measuring error.

If the current must be measured without opening the circuit under test, a current probe needs to be used. This is placed around the wire or cable through which the measurand flows. Its output is supplied to an ammeter or oscilloscope. Such probes are suitable only for measuring medium to high currents. Probes that operate on the transformer principle are suitable only for measuring alternating currents, whereas those that use a Hall generator may be used for d.c. and a.c. measurement.

Some tips

Except when current is measured with a probe, the circuit under test should be switched off when it is being prepared for the test. Only when the measuring instrument is secured to it should the power be switched on. Also, make sure that no large charging currents (caused by uncharged electrolytic capacitors, for instance) can overload the instrument or blow its fuses.

When the approximate value of the measurand is not known beforehand, the highest metering range of the instrument used should be selected. The most appropriate range may then be selected once the measurement is underway.

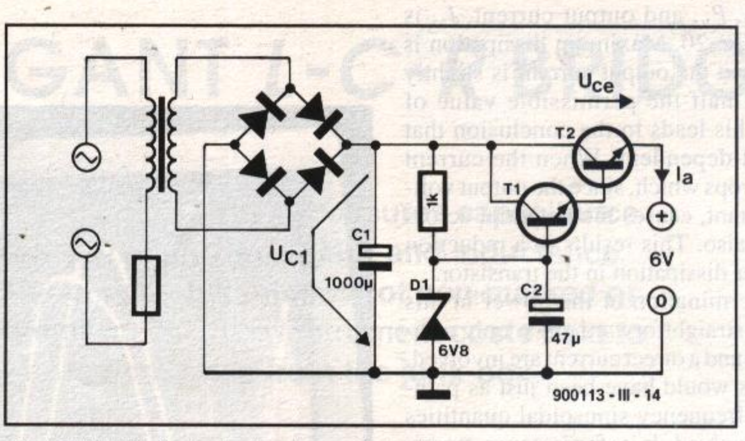

Fig. 19. Circuit of a typical series regulator in a power supply. The example in the text is the same when an integrated voltage regulator is used.

Measurement of power

Special power meters with separate current and voltage circuits are hardly ever necessary. In most cases, it is perfectly all right to measure current and voltage separately in the circuit under test and then calculate the power dissipated.

A typical example is the dissipation in a series regulator—see Fig. 19. In this circuit, a regulated voltage of 6 V is obtained from the unregulated potential across C_1. The dissipation in T_2, which is the product of voltage U_{ce} across T_2 and output current I_a, is converted into heat.

Of interest is the operating point at which the dissipation in the transistor is a maximum. That point depends to some extent on the voltage across C_1. The relation between power

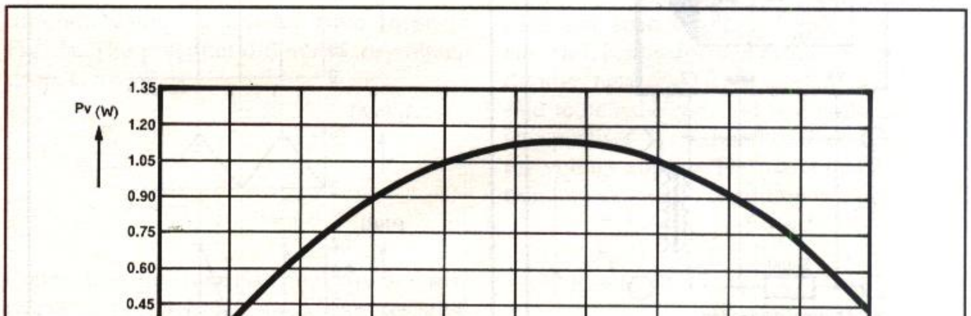

Fig. 20. Curve showing power dissipation vs output current in a typical series regulator.

dissipation, P_V, and output current, I_a, is shown in Fig. 20. Maximum dissipation is reached when the output current is slightly larger than half the permissible value of 330 mA. This leads to the conclusion that U_{c1} is load-dependent. When the current rises, U_{c1} drops which, since the output voltage is constant, causes the potential across T_2 to drop also. This results in a reduction in the power dissipation in the transistor.

The determination of the power in this example is straightforward since only a direct voltage and a direct current are involved. The process would have been just as painless if low-frequency sinusoidal quantities had been involved. It is, for instance, possible to calculate the input power of the transformer in Fig. 19 once the voltage and current are known. A word of caution here, though: since the alternating measurands are operating with a reactance, that is, the transformer winding, the calculation will result in the **apparent power**. This is not of such practical use as the **active** (or **true**) **power**. To determine that, the power factor cosϕ must be known. The true power, P, is then calculated from

$$P = U \times I \times \cos\phi.$$

The power factor may be determined with the aid of a dual-trace oscilloscope as shown in Fig. 21. One trace shows the voltage and the other the current. Not the magnitude of the two measurands is important, but the phase difference between them: this is the distance between the zero crossing of the voltage and that of the current. Unfortunately, the current in Fig. 21 is measured inverted; this cannot be avoided, however, and must be allowed for in the calculation. A number of scopes have an invert mode facility, which enables inversion of the trace on the screen.

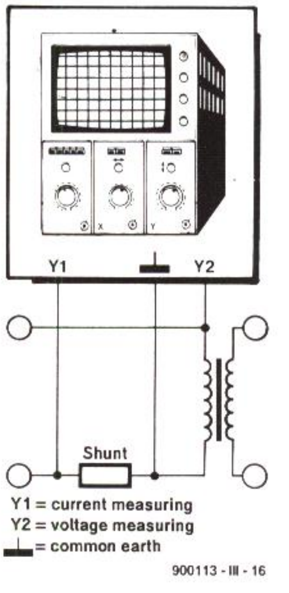

Fig. 21. Phase shift may be measured with the aid of a dual-trace oscilloscope. One signal will be inverted since there is only one common earth.

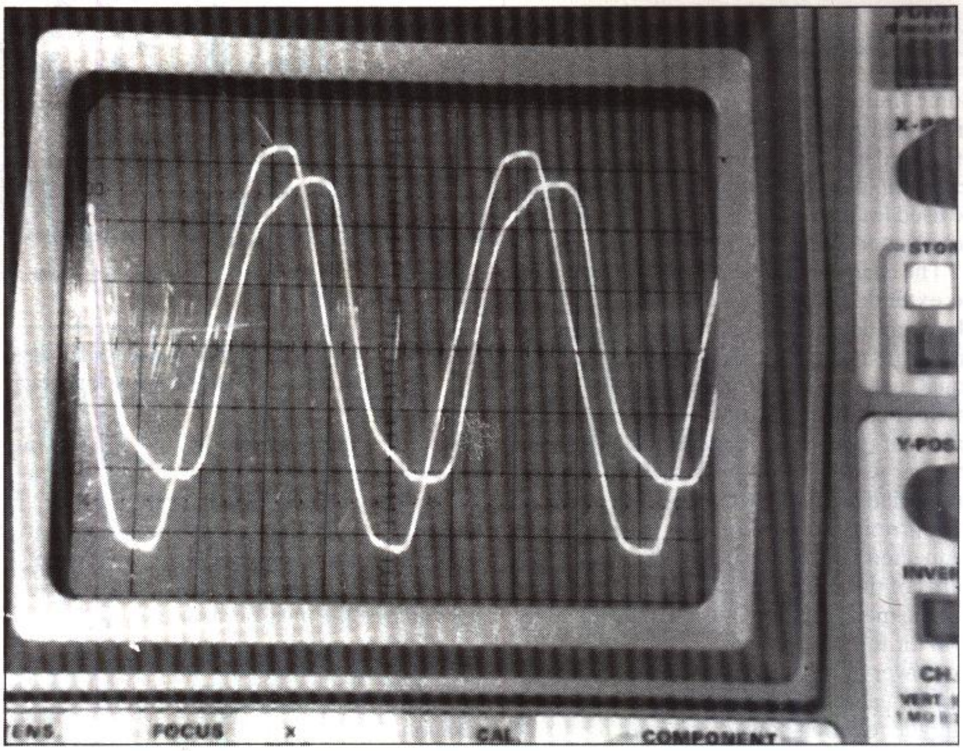

Fig. 22. Measurement of phase shift between voltage and current pertaining to a soldering iron operated from the mains via an isolating transformer.

Such a screen is shown in Fig. 22: the two traces pertain to the voltage and current of a soldering iron operated from the mains via an isolating transformer. The value of the shunt is 1 Ω; the error caused by it may be ignored. The phase shift between the two traces is 2 scale divisions. Since a period of 360° extends over 20 divisions, the phase shift, ϕ, amounts to

$$\phi = (360 / 20) \times 2 = 36°,$$

and the power factor is

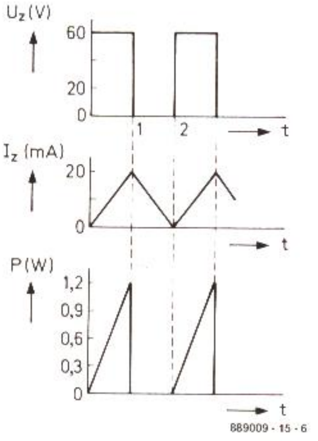

Fig. 23. A dual-trace oscilloscope may also be used to determine the power when voltage and current are not sinuoidal.

$$\cos\phi = \cos 36° = 0.809.$$

Such measurements on the mains must be made with the aid of an isolating transformer, since otherwise a short-circuit may be caused by the earth of the oscilloscope.

The relation between the input power and output power of a transformer is the efficiency, ε. To get a correct analysis of the behaviour of a mains-operated power supply, some additional calculations are necessary. These types of measurement become easier to handle when the subsequent computation is carried out by a computer and relevant software. A suitable table may be designed with the aid of a table calculation program. Constants and formulas may be taken from memory, so that only the actual measurements need to be entered. The results, that is, input power, output power, power dissipation, efficiency, and so on, then become available in tabular form, from which relevant curves or bar diagrams may be produced alsmost immediately the measurements have been taken. A printer may be be found useful to put all the information to paper.

Power measurements become a little more difficult when switching regulators or phase gating circuits are involved. The waveform is then not sinuoidal, so that most multimeters do not give a correct result: a true RMS meter is then required. The two waveforms may also be displayed on the screen of a dual-trace oscilloscope for analysis. For a worthwhile result, it is best to redraw the curves on suitable graph paper, as for instance in Fig. 23, to arrive at a power curve. The average value deduced from that curve is the required power. ■

A SIMPLY ELEGANT *L-C-R* BRIDGE

The balanced bridge described in this article measures capacitance from 1 pF to 10 μF, resistance from 1 Ω to 10 MΩ, and inductance from 1 μH to 100 μH. Ideal for checking the values of non-marked or otherwise non-identifiable components, the instrument costs next to nothing, and can be built from parts from the junk-box.

E. Chicken, MBE, MSc, BSc, CEng, FIEE (G3BIK)

THE first design of the instrument proposed here was a basic capacitance-measuring bridge which consisted of nothing more than two capacitors, a 10-kΩ carbon linear potentiometer, a crystal earpiece, and a total of four solder joints, as shown in Fig. 1.

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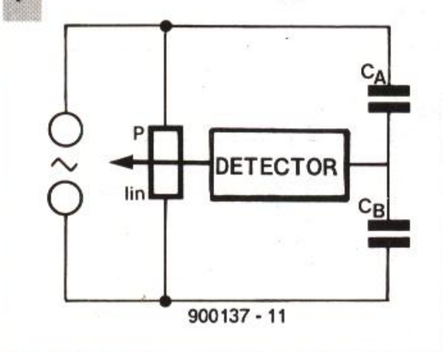


An alternating voltage at audio frequency was required to feed the bridge, and a quick glance around the author's radio shack showed a number of possible sources, e.g., an oscilloscope which offered a 1-kHz square-wave signal at 1 V_{pp} on its front panel for self-calibration, an RF signal generator which offered a 1-kHz sine-wave at a few volts rms, a home-made AF oscillator with a 2-V output, and, if the worst came to the worst, the audio output from the transistor broadcast radio tuned to the pop-music channel!

All were tried and found to provide a fully audible signal in the crystal earpiece, and rotation of the balance potentiometer spindle yielded a clearly discernible and sharp null, more than adequate to afford repeatable accuracy of measurement.

The capacitors were replaced with resistors, with equally promising results for the measurement of resistance.

But what of inductance in the RF range? A few turns of enamelled copper wire were wound on to an available 7-mm former with an iron-dust slug core. A 100-pF capacitor was soldered across the coil. Its resonant frequency was adjusted to about 15 MHz with the aid of a gate dip oscillator. Next, the inductance of the coil was calculated to be approximately 1 μH. That would be the 'unknown' inductor of low value.

Another coil was wound with about ten times as many turns to produce a higher value of inductance. The actual value was not important, as will be explained later.

With the two inductors connected into the bridge in place of the two capacitors, again it was possible to obtain a sharp audio-null to indicate balance.

Turning the core in the 1-μH coil to produce a different value of 'unknown' inductance required the potentiometer to be re-adjusted to restore balance. This demonstrated the viability of the simple bridge as a measuring device suitable for all three types of passive component, i.e., for inductance (*L*), capacitance (*C*) and resistance (*R*).

Balanced bridge for *L-C-R* measurement

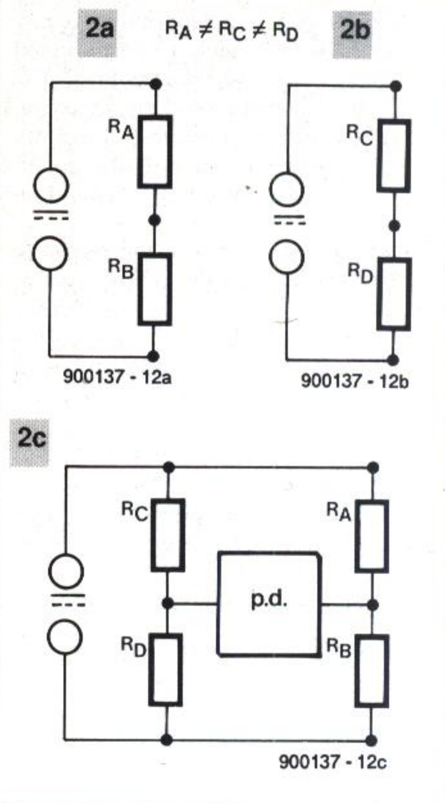
Consider first the simple resistive potential divider shown in Fig. 2a. The potential difference, or voltage drop, across R_a is

$$U_s \frac{R_a}{R_a + R_b}$$

where U_s is the supply voltage. Now consider the other resistive potential divider in Fig. 2b, which incorporates unspecified but different values of resistor from those in Fig. 2a. The potential difference, or voltage drop, across R_c is

$$U_s \frac{R_c}{R_c + R_d}$$

Connecting the two potential divider networks in parallel to form a basic resistive bridge, and feeding them both from the common voltage supply, as shown in Fig. 2c, does not alter the two equations given for the potential differences across R_a and R_c.



If, however, the voltage drop across R_a is different from that across R_c, then a voltage detector connected between junction R_a-R_b and junction R_c-R_d will indicate the difference between the two voltage-drop values. But, if the voltage drop across R_a is equal to that across R_c, the potential difference will be nought, and the voltage indicator will read zero or a voltage 'null'. The electrical bridge formed by the two potential divider networks R_a-R_b and R_c-R_d is then said to be balanced, and that will also be so irrespective of whether d.c. or a.c. is used for the voltage supply. The mathematical equation for such a null condition is given by

$$U_s \frac{R_a}{R_a + R_b} = U_s \frac{R_c}{R_c + R_d}$$

which simplifies to

$$\frac{R_a}{R_b} = \frac{R_c}{R_d}$$

If the value of R_b is known and fixed, and if the ratio R_c/R_d is known, it is possible under this null condition (i.e., when the bridge is at balance) to determine the actual value of an unknown resistor R_a from

$$R_a = R_b\frac{R_c}{R_d} = (\text{known } R_b)\times(\text{ratio } R_c/R_d)$$

For example, if R_c is made equal to R_d, their ratio is 1.0, and the null condition becomes R_a=R_b, irrespective of the actual values of Rc and R_d. Only their ratio is of importance.

Further, if the ratio R_c/R_d is made adjustable and calibrated, and if the value of the known resistor R_b is made to be switch-selectable, the unknown resistor, R_a, can be quantified over a wide range of values.

Scale design: the basics

For convenience, R_c and R_d can be replaced by a linear-law rotary potentiometer of any convenient resistive value, to be provided with a circular scale and a pointer knob. The scale is then calibrated to read the ratio of resistive values measured between the centre tag and the two outer tags of the potentiometer, for different angles of spindle-rotation.

At mid-travel, for example, the resistance measurements between the centre tag and the two outer tags would be equal in an ideal linear-law potentiometer. At three-quarters traverse, the resistance of one section would be some three times that of the other section, giving a ratio of 3:1 and vice-versa.

In other words, as the spindle is rotated away from mid-position in one direction, the ratio will increase from 1.0 upwards towards infinity, and in the other direction it will decrease from 1.0 towards zero.

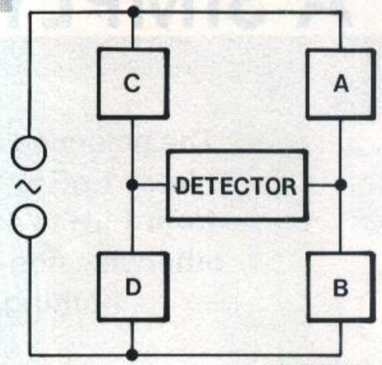
Although the rotational movement of a standard linear-law carbon potentiometer is restricted to about 300 degrees of travel, fortuitously and very conveniently the ratios ×10 and ×0.1 fall at approximately 90 degrees on either side of mid-travel. So, if mid-travel is positioned at the top of the scale, i.e., at 360 degrees, then the dial can be marked ×0.1, ×1.0, and ×10 at 270, 360 and 90 degrees respectively, and ×0.01 and ×100 at 240 and 120 degrees respectively; to a first approximation.

The value of a resistor is thus readily determined by rotating the potentiometer until a null is detected. The ratio indicated on the scale is subsequently multiplied by the value of the known resistor, R_b. Mathematically,

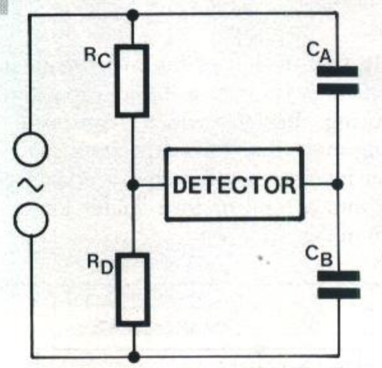
$$R_a = R_b \times [\text{ ratio } R_c/R_d \text{ at null }]$$

By assigning a selection of different values for the *known* resistor R_b, the range of measurement for the *unknown* resistor, R_a, can be conveniently modified. To simplify measurement even further, it is preferable to use whole number values for the *known* resistors, e.g., 100 Ω, 10 kΩ, 100 kΩ, etc. These known resistors, which for convenience can be range-selectable, are then renamed *range* resistors, and the switch escutcheon is marked with their values. An example: if a null occurs at, say, ×0.05 on the 1,000 Ω range, the unknown resistor has a value of (1,000 × 0.05) = 50 Ω.

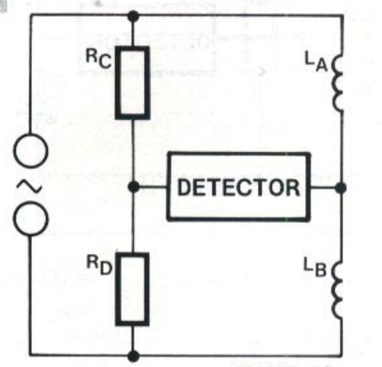
3a



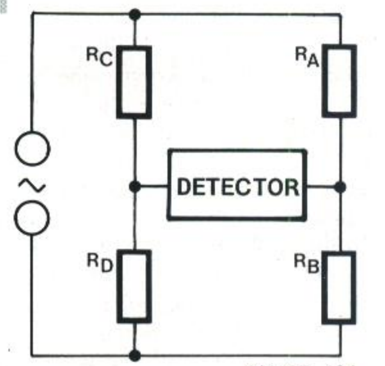
3b



3c



3d



Inductance measurement

The above principle of measurement can also be applied to inductance, with resistors R_a and R_b replaced by inductors L_a and L_b. To be effective, the bridge *must* be fed with an a.c. voltage at a frequency, *f*, high enough to produce inductive reactances X_a and X_c respectively, sufficient to provide potential differences suitable for null detection.

The simplified potential difference equations at null condition are:

$$X_a = X_b\frac{R_c}{R_d}$$

where X for inductive reactance is

$$X = 2\,\pi\, f\, L$$

So, this equation simplifies to:

$$L_a = L_b\frac{R_c}{R_d}$$

which conveniently uses the same ratio multiplier R_c/R_d as for resistance measurement, hence the same ratio scale can serve for both resistance and inductance.

Capacitance measurement

For the measurement of capacitance with inductors L_a and L_b replaced by capacitors C_a and C_b, again the voltage supply must be a.c. to provide, in this case, capacitive reactances X_a and X_b.

The simplified equation for potential difference in the null condition is

$$\frac{X_a}{X_b} = \frac{R_c}{R_d}$$

where X for capacitive reactance is

$$X = \frac{1}{2\,\pi\, f C}$$

$$X_a = X_b\frac{R_c}{R_d}$$

which simplifies to

$$C_a = C_b\frac{R_d}{R_c}$$

Note that in this expression the ratio multiplier, R_d/R_c, is inverted with respect to the one used for the resistance and inductance measurements. In practice, this means that a mirror-image of the ratio scale is required for the measurement of capacitance. So, the scale

0.01 0.1 1.0 10.0 100.0

becomes

100.0 10.0 1.0 0.1 0.01, etc.

Again, a selection of range capacitor values for C_b will provide the desired range of measurement, and for user convenience,

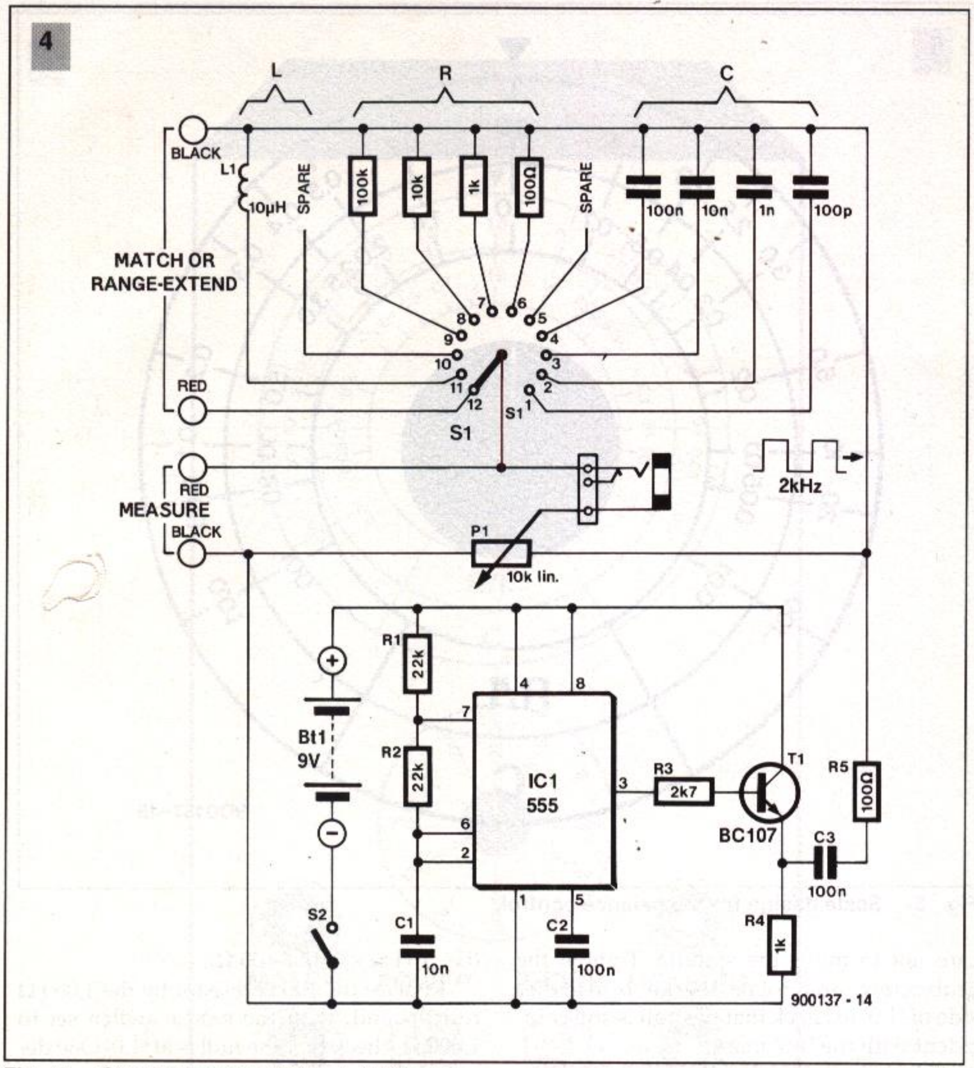


Fig. 4. Circuit diagram of the *L-C-R* meter.

they should be round numbers, e.g., 100 pF, 0.1 µF, etc.

Circuit design concept

The circuit diagram of the *L-C-R* bridge is given in Fig. 4. A rectangular wave of about 9 V_{pp} is provided by a 555, IC1, configured as an astable multivibrator, with a repetition frequency of about 2 kHz.

The 2-kHz output from the astable is buffered by an emitter follower transistor, T1, to minimize loading of the 555 output circuit by the bridge when this is switched to the lower resistance/reactance ranges.

A crystal earpiece is used for the null detector. Its high impedance offers a better audible signal than would an electromagnetic version to help with the determination of the null, particularly when measuring inductance.

Accuracy of measurement depends mainly on the accuracy of the component values switched into the circuit by the *L-C-R* range switch, S1, and on the quality of the linear-law potentiometer, P1, used for the balance control.

The prototype of the *L-C-R* bridge was built with a standard off-the-shelf linear carbon potentiometer for the balance control (a 1,000-Ω version should serve equally well), low-tolerance, high-stability resistors and capacitors, and a commercially available moulded RF inductor of 10% tolerance.

It was considered that a single 10-µH range inductance would provide a wide enough range of inductance measurement for normal RF purposes, but provision has been made for extending the measuring range for each type of component. This can be provided permanently by wiring an additional L_b, C_b, or R_b on to the range switch, or temporarily by connecting an appropriate component across the MATCH terminals with the RANGE switch in the 'match' position.

The MATCH terminals serve also to allow value matching of a pair of external components. When the two are exactly equal in value, the null falls exactly at the '1.0' position on the ratio scale, and when not matched, the scale indicates the relative value of the component connected to the MEASURE terminals compared to that at the MATCH terminals.

Although no d.c. polarizing voltage has been included for electrolytic or tantalum capacitors, the bridge can be used for the measurement of such capacitors, without problems.

Current drain from the 9-V PP3 battery is only about 7 mA.

Construction

Because electrical shielding is not required, the components are housed in a low-cost plastic container with removable flat cover. Externally accessible components, i.e., balance potentiometer, range switch, MATCH and MEASURE terminals, ear-piece socket and battery switch, are mounted on to the removable panel.

One lead of each of the 'range' components is soldered directly to the appropriate lag-terminal of the RANGE switch, and the other lead of each component is soldered to a self-supporting ring of tinned copper wire.

The RANGE switch is a single-pole 12-position rotary midget wafer type, with a plastic spindle and fixing bush, to minimize stray capacitance which might adversely affect the measurement at the lower picofarad range. A flat on the spindle allows a push-on knob to be used, of the type which has a moveable cover-cap to allow alignment of the pointer.

COMPONENTS LIST

Resistors:

2	100Ω 1% metal film	1 for range switch, 1 for calibration
2	1kΩ 1% metal film	1 for range switch, 1 for calibration
2	10kΩ 1%, metal film	1 for range switch, 1 for calibration
2	100kΩ 1%, metal film	1 for range switch, 1 for calibration
2	22kΩ	R1;R2
1	2kΩ7	R3
1	1kΩ	R4
1	100Ω	R5
1	10Ω linear potentiometer	P1

Capacitors:

2	100pF 1% silvered mica	1 for range switch, 1 for calibration
2	1nF 1% silvered mica	1 for range switch, 1 for calibration
2	10nF 1% polystyrene	1 for range switch, 1 for calibration
2	100nF 5% polyester	1 for range switch, 1 for calibration
2	10nF ceramic	C1;C2
1	100nF ceramic	C3

Inductors:

1	10µH encapsulated polypropylene
1	10µH air-cored, self-wound (see text)

Semiconductors:

1	555	IC1
1	BC107	T1

Miscellaneous:

1	enclosure, ABS plastic, 120×65×40 mm, with lid
1	1-pole, 12-way rotary midget wafer switch, plastic spindle/bush
1	miniature on/off switch
1	3.5-mm panel mounting jack socket
1	crystal ear-piece
1	DIL socket 8-way
1	35-mm square section of 0.1-inch hole-spacing copper-strip board
2	4-mm terminal post, red
2	4-mm terminal post, black
1	PP3 battery with twin press stud and holder

The drawing in Fig. 6 shows a suitable escutcheon for the L-C-R SELECT/RANGE switch.

The linear-law carbon potentiometer used for the balance control also has a plastic spindle and fixing bush, but preferably without a flat on the spindle, to allow a grub-screw type of pointer knob to be aligned to the '1.0' mark on the ratio-scale.

The 555 integrated circuit, transistor T1, and the eight associated passive components are assembled on to a 35-mm square piece of 0.1-inch hole-spacing copper strip board. It is reasonable to use an 8-way DIL socket for the 555. The actual lay-out of the few components on the board is uncritical, hence does not warrant a guidance sketch. The finished 2-kHz oscillator board is small enough to be self-suspended by its connecting wires, but its PP3 battery may need an elementary fixing bracket or zip-strap.

Four 4-mm terminal posts for the MEASURE and MATCH pairs of terminals allow the components to be either loosely plugged in, or more securely screw-fastened to suit the circumstance.

The size of the enclosure is not critical, indeed the prototype used a plastic box of about the same general shape and size as a standard wall-socket box of 75×75×45 mm, but a box with a panel of, say, 65×120 mm would accommodate a circular scale of readable dimension plus the range switch escutcheon, with room to spare for the terminals, battery switch, and ear-piece socket.

Calibration

The ratio scale needs to be calibrated for optimum results, because the linear-resistance characteristic of the chosen balance potentiometer may not necessarily be the same as that used in the author's prototype. The sample scale shown in Fig. 5 may be used for guidance.

Although the standard potentiometer has a rotational travel of about 300 degrees, it is recommended that the usable ratio scale be confined to ±90 degrees about centre (ratios: ×0.1, ×1.0, ×10), or at most ±120 degrees (ratios ×0.01, ×1.0, ×100).

The easiest way to calibrate the scale is to remove the pointer knob and temporarily affix a circular paper-scale centrally over the fixing bush, with the '1.0' mark uppermost at 360 degrees.

The circular scale should have two circular bands, one marked 'R/L' for resistance and inductance, and the other marked 'C' for capacitance.

Plug in the ear-piece, select the 100 Ω range, and switch on the oscillator. The parts list recommends the availability of four duplicate low-tolerance high-stability resistors. These will be used as the external standards for calibration/test of the scale.

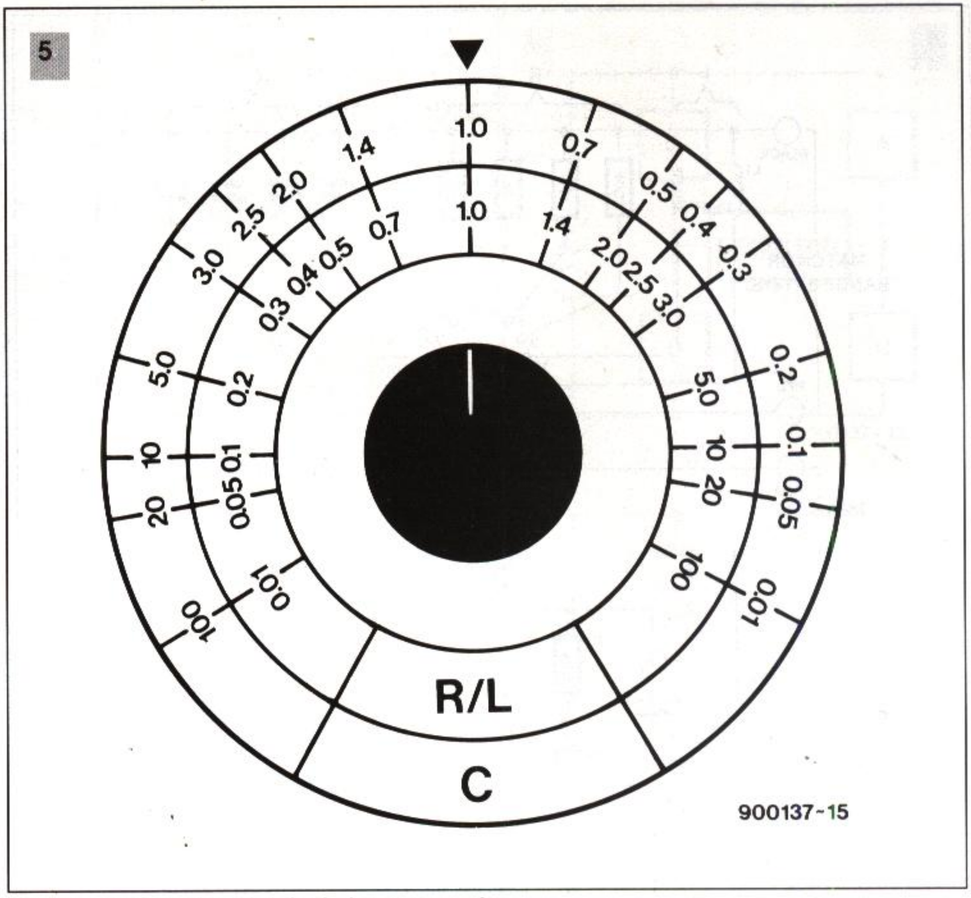


Fig. 5. Scale design for the balance control.

Connect the 100 Ω resistor to the MEASURE terminals, and carefully rotate the balance pot spindle until a null is obtained in the ear-piece. Loosen the grub-screw in the pointer-knob, and carefully position the knob on to the spindle with its pointer exactly at '1.0' (360 degrees) position of the scale, taking care not to move the spindle. Tighten the grub-screw, and rotate the knob to either side of '1.0' to check that the null is still coincident with the '1.0' mark.

Select the 1,000 Ω range and, with the 100 Ω resister still connected, adjust the pointer for null in the ear-piece. This should occur at about 270 degrees. Mark the R/L scale '×0.1' at that position (i.e., 1,000 Ω × 0.1 = 100 Ω).

Switch to the 10 kΩ range, when the null should appear at about 240 degrees, and mark the R/L scale '×0.01' at that position (i.e., 10 kΩ × 0.01 = 100 Ω).

Replace the 100 Ω resistor by the 1,000 Ω resistor and, with the RANGE switch set to 1,000 Ω, check that the null is at '1.0' (360 degrees).

Switch to the 100 Ω range, still with the 1,000 Ω resistor, when the null should occur at about 90 degrees. Mark the R/L scale '×10' at that position (i.e., 100 Ω × 10 = 1,000 Ω).

Replace the 1,000 Ω resistor by the 100 kΩ resistor, set the range switch to 1,000 Ω, and the null should occur at about 120 degrees, and mark the R/L scale '×100' at that posi-

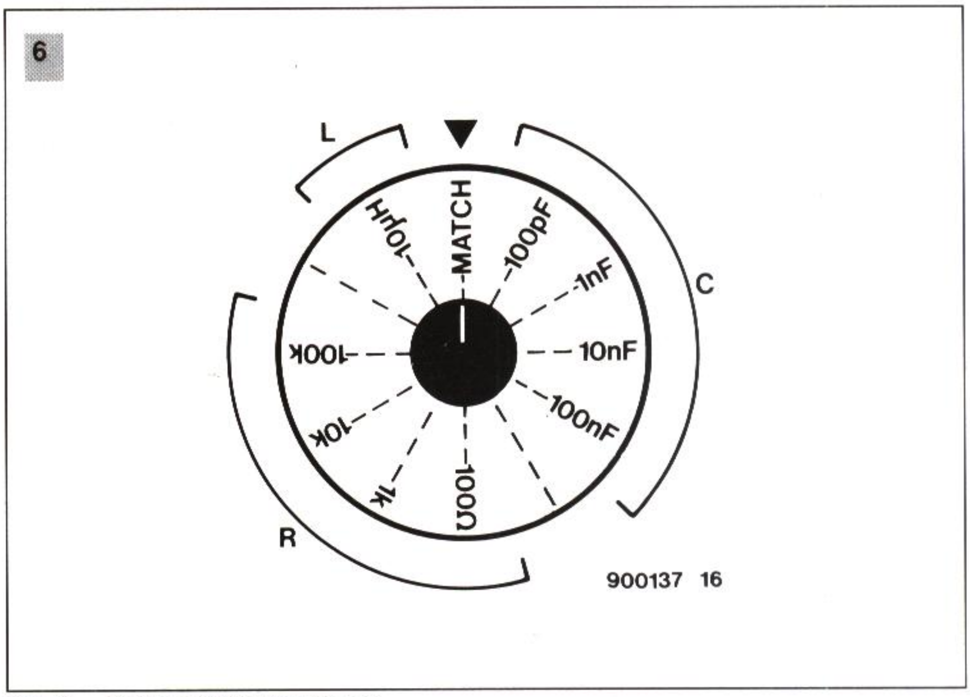


Fig. 6. Range switch escutcheon.

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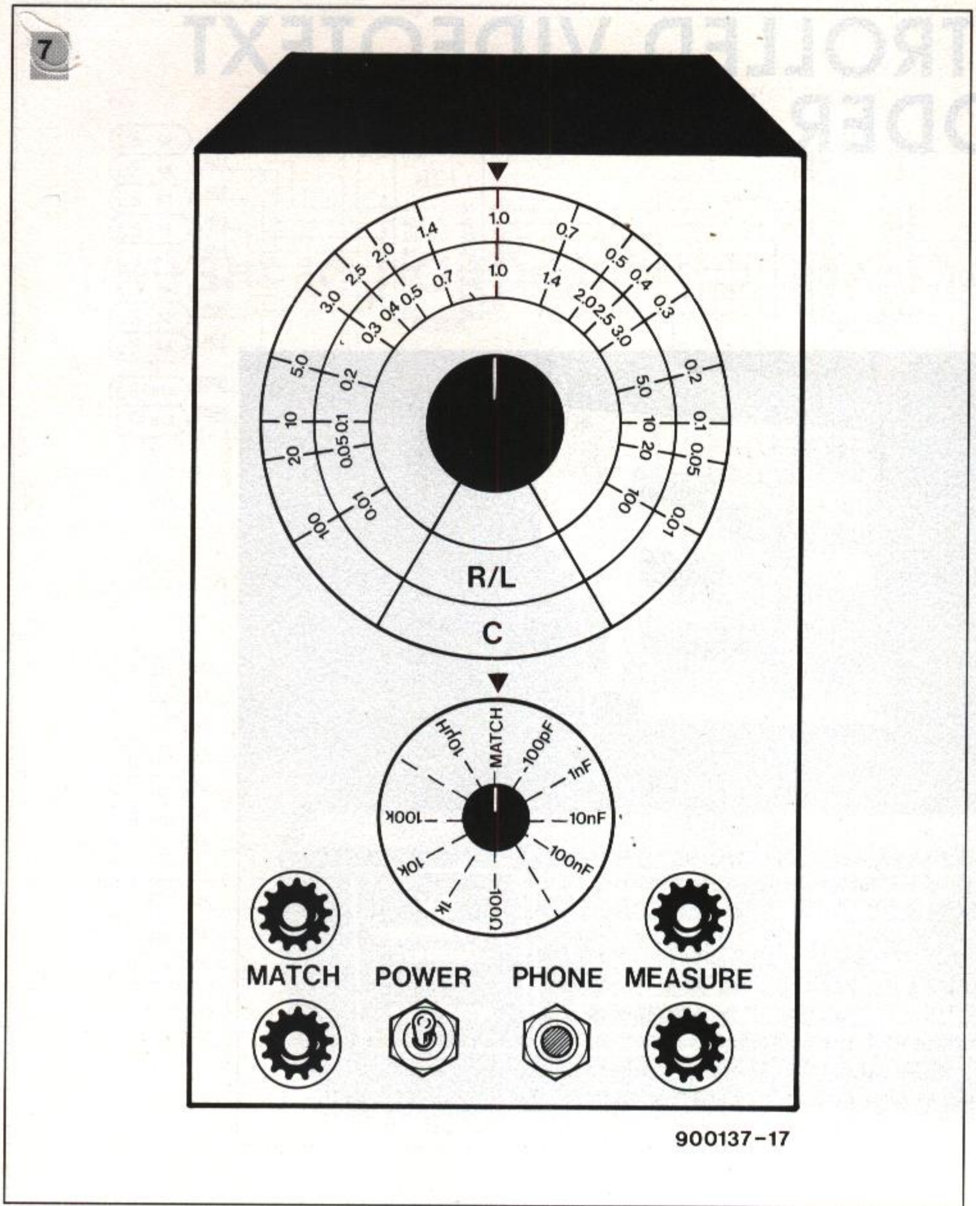


Fig. 7. Suggested front panel layout of the *L-C-R* bridge.

R/L marking	C marking
0.1	10
0.2	5.0
0.3	3.3
0.4	2.5
0.5	2.0
0.6	1.7
0.7	1.4
0.8	1.3
0.9	1.1
1.0	1.0
0.01	100
0.02	50
20	0.05
80	0.013

tion (i.e., 1,000 Ω × 100 = 100 kΩ).

It will now be obvious that the ratios ×0.1 to ×1.0 and ×1.0 to ×10 each span about 90 degrees of travel, whereas the ratios ×0.01 to ×0.1, and ×10 to ×100, each span only about 30 degrees of travel. This means that calibration points beyond less than ×0.1, and greater than ×10, become increasingly cramped — but still very repeatable and acceptably accurate provided that care is taken with the calibration of the intermediate points in each sector.

Nulls are still very detectable even further towards the end-stops of travel, but the cramping is even more pronounced. That is where the advantage of a large diameter scale becomes apparent. The majority of *in vivo* measurements will however fall within the ±90 degree bands, where the scale is uncramped and clearly readable.

To calibrate the intermediate scale-points, it is advisable to restrict them to the 120 degree sectors on either side of '1.0', and to use whole numbers for the external calibration resistors rather than the decimal values of the 'preferred' series. The easiest way to do this is to use a 1,000 Ω standard linear-law potentiometer with a wire connected to its centre tag, and another to one of its outer tags, and to set it to a selection of values by means of an ohm-meter, i.e., 200, 300, 400 … 900 Ω, etc. These values are then used to establish the intermediate scale points, i.e., ×0.02, ×0.03, ×0.04 … ×0.09 using the 10 kΩ range; ×0.2, ×0.3, ×0.4 … ×0.9 on the 1,000 Ω range; and ×2, ×3, ×4 … ×9 using the 100 Ω range.

Similarly with values between 20 kΩ and 90 kΩ on a 100 kΩ potentiometer, to give scale points of ×20, ×30, ×40 … ×90 using the 1,000 Ω range.

The resistance calibration scale applies equally to the measurement of inductance, but a mirror-image scale is required for capacitance measurement. The same calibration positions pertain, but the C scale must be marked with the inverse values from the R/L scale points, rounded up for practical purposes, i.e., as shown in the table. Calibration is now complete, and the bridge is ready for use.

Does it work?

Yes. Results when measuring resistance can be accepted with confidence, as can measurements of capacitance, provided that the capacitors being measured are of good electrical quality. However, regarding the measurement of inductance, it must be borne in mind that the balance equations have been simplified by assuming zero resistance in the small values of inductance to be measured.

In practice, this will not be the case, depending on the construction of the particular RF coil. For example, the moulded 10 µH inductor used in the prototype has a series resistance of about 0.3 Ω owing to the very thin wire used for the coil, and whilst this does not unduly detract from the calibration accuracy when measuring inductors wound with similarly thin wire gauges, it does create inaccuracies when trying to measure the values of coils with heavier gauge wires, e.g., 14 to 24 SWG (14 to 25 AWG).

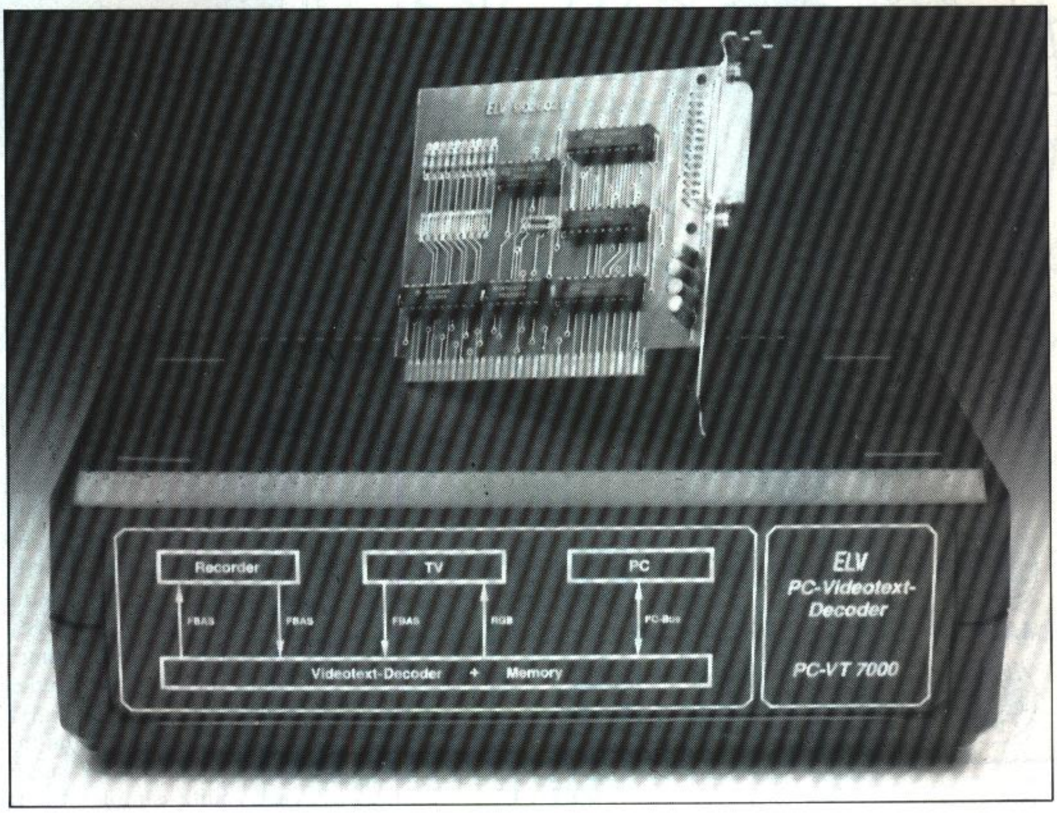
The solution to the measurement of inductance of heavier gauge coils is nevertheless quite simple, by using the MATCH terminals with an alternative 'standard' 10 µH inductor connected across them, but wound from thickish wire. A suitable alternative 10 µH air-cored coil can be constructed easily by close-winding 24 turns of 20 SWG (1.0 mm dia., or 21 AWG) enamelled copper wire with a winding-span of about 24 mm on to a PVC former of 25 mm outside diameter; or 32-turns of 20 SWG wire by 32 mm span on to a 20 mm outside o.d. PVC former; or 69 turns by 69 mm span of 20 SWG wire on to a 12.5 mm former. In each case, allow 10 mm end-tails.

Measurement of thicker-wire unknown inductance now follows normal procedure, but with the switch set at its 'match' position instead of the 'L' position, and with the 10 µH thick-wire standard inductor connected to the MATCH terminals.

To get the feel of the bridge, try measuring a selection of *L*, *C* and *R* components, and the effect of tolerance on nominal values. Then try matching component values by connecting pairs of nominally equal values to the MATCH and MEASURE terminals until null is obtained at '1.0' on the scale. Try also extending/modifying the range of measurement by connecting a known value component to the external MATCH terminals. And for interest only, try measuring the relative effect of an iron-dust slug versus a brass slug in a low-value RF inductance. ■

PC-CONTROLLED VIDEOTEXT DECODER PC-VT7000

PART 2: I^{2}C INTERFACE AND CONSTRUCTION OF THE UNIT



This second and concluding part of the article deals with the operation of the I^{2}C PC insertion card that forms the link between the computer and the main decoder. As usual, we close off the article with full constructional details of the project.

THE term I^{2}C refers to a control and data bus system developed by Philips Components for use on ICs in consumer electronics equipment (Ref. 2). The I^{2}C system is used here to control the SDA5243 ECCT on the main decoder board (see Part 1 of this article). A special PC insertion card has been developed to ensure the fastest possible two-way communication between the ECCT and the PC, with the aid of the I^{2}C bus.

The block diagram of the I^{2}C bus controller card is shown in Fig. 6. At the side of the PC extension slot, an address comparator compares the I/O addresses supplied by the PC with a user-defined address. If the addresses match, the comparator supplies an appropriate enable signal to a 'control logic' block, which, in turn, switches on two databus buffers and an 8-bit D-latch. When the card is written to with an appropriate I/O address, data on the PC databus is copied to an 8-bit D-latch. The open-collector drivers

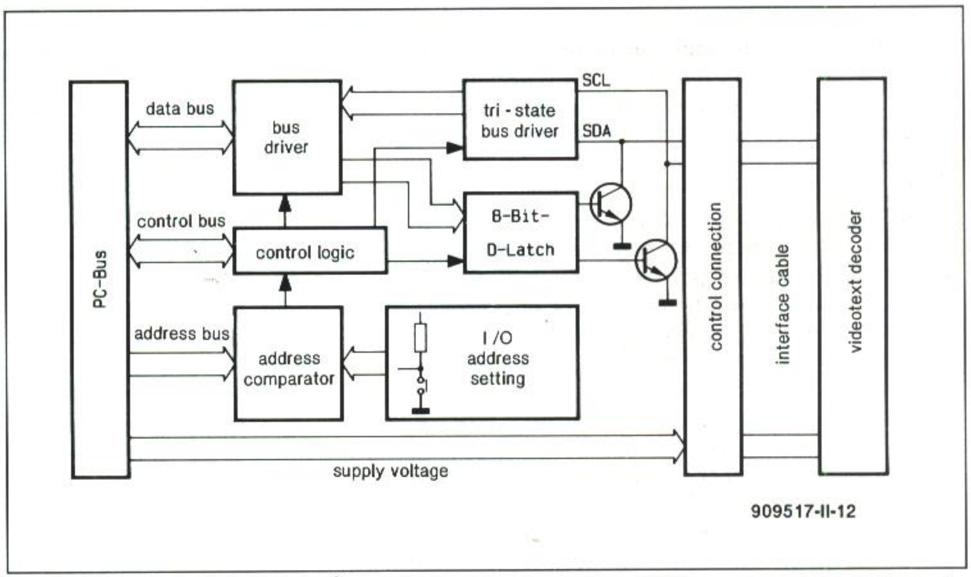


Fig. 6. Block diagram of the I^{2}C card. This forms an interface between the PC and the main decoder, which is a separate unit in this project.

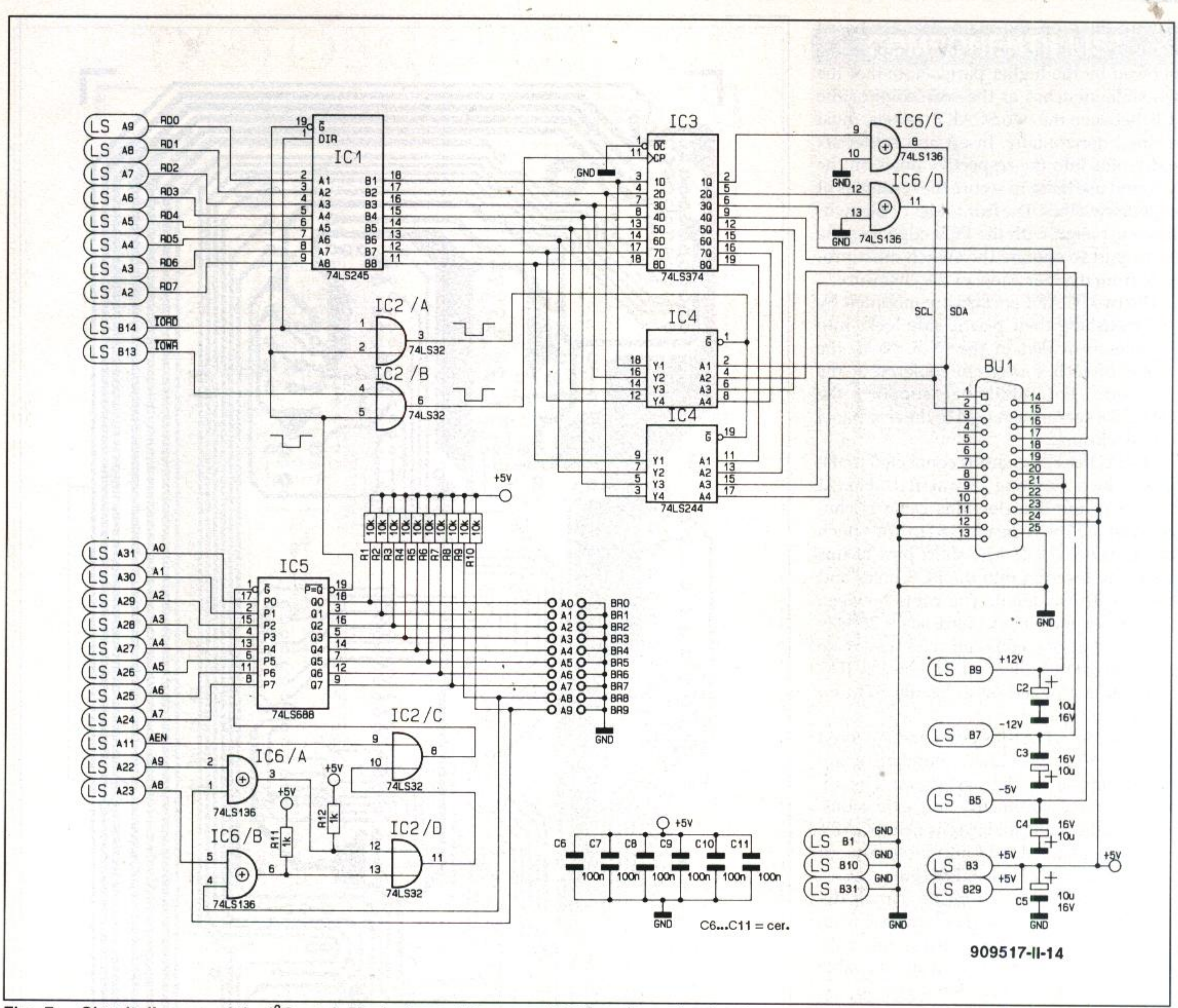


Fig. 7. Circuit diagram of the I^2C card. Basically an I/O mapped PC interface, the circuit provides the bidirectional SCL and SDA lines.

at the outputs of the D-latch provide the transmit function of the insertion card for two I^2C lines, SCL (serial clock) and SDA (serial data). Since these lines are bidirectional, provision is also made to convey information from the videotext decoder to the I^2C card. A 3-state busdriver is enabled when the PC reads from the card address. The SCL and SDA information captured by the 3-state busdriver is fed to the PC via the databus buffer.

Circuit description of the I^2C card

Figure 7 shows the detailed circuit diagram of the PC-compatible I^2C interface card. The bidirectional databus buffer is formed by IC1, a 74LS245. The data direction is switched under the control of PC bus line $\overline{\text{IORD}}$ (pin B14), which is also connected to IC2A. The second PC control line, $\overline{\text{IOWR}}$ (pin 13) is connected to another gate, IC2B. The two OR gates IC2A and IC2B form the 'control logic' block shown in Fig. 6. Their function is enabled by the address comparator when the card is addressed by the PC.

The address of the card in the PC's I/O map is defined by wire jumpers BR0 to BR9. A '0' is set for a particular address bit by fitting the associated wire. The address comparator, IC5-IC6A, is not enabled until PC bus line AEN is logic high, while at the same time either $\overline{\text{IORD}}$ or $\overline{\text{IOWR}}$ is logic low. If this condition is satisfied, and if the card is addressed by the PC, the output of IC5, pin 19, changes from high to low, thus enabling the rest of the circuit via IC2A-IC2B.

The logic level on the $\overline{\text{IORD}}$ and $\overline{\text{IOWR}}$ lines determines whether the PC writes to 8-bit D-latch IC3 (a 74LS374), or reads from 3-state bus driver IC4 (a 74LS244).

Outputs 1Q and 2Q of latch IC3 are connected to the inputs of XOR gates IC6C and IC6D. These LS-TTL gates have open-collector outputs, and function as drivers for the I^2C lines, SCL and SDA. They have no logic function, and are equivalent to the two transistors shown in Fig. 6. SCL and SDA information received from the videotext decoder reaches the I^2C card via inputs A1 and A4 of 3-state bus driver IC4.

The signals on outputs 5Q to 8Q of IC3 are fed back to the inputs of the IC via buffer IC4. This allows the card to be detected by the control software, and an error message to be generated when it is not fitted or not found at the requested address.

Outputs 3Q and 4Q of IC3, and inputs A2 and A3 of IC4, are taken to a 25-way D-connector, BU1, for future extensions. The 25-way connector also serves to connect the approximately 2-m-long cable to the videotext decoder. Apart from data and commands, this cable also carries the +12 V, –12 V, –5 V and +5 V supply voltages, and a common ground line, from the PC to the decoder board.

Construction

Briefly recapitulating, the PC-VT7000 consists of two units: the main decoder, which is housed in a black, 7000-series ABS enclosure, and the I^2C card, which is inserted into a free extension slot in your PC. These two units are linked by a cable.

Decoder board

The construction is best started by fitting all

35 wire links on the main decoder board (Fig. 8). Next, fit the low-profile components, followed by the higher parts. Note that the two slide switches at the rear edge of the PCB, between the two SCART sockets, must be fitted horizontally. Insert and solder six solder pins into the respective holes on the PCB, and use these to secure the terminals of the slide switches. The front sides of the switches must align with the PCB edge, i.e., the plastic part to operate the switch must protrude from the rear panel of the enclosure.

The two SCART sockets are mounted by gently pushing their plastic side locks into the holes provided in the PCB. Next, the 21 pins of each socket are soldered at the track side. For additional support, the SCART sockets are screwed to the rear panel of the enclosure.

The PC insertion card is connected to the decoder by a 2-m-long 25-way flatcable. At the side of the decoder, this cable is connected to a 26-way header, STL1 (of which one pin is not used). The short pins of this header are inserted into the PCB holes and subsequently soldered. The cable between the decoder and the I^{2}C card has a 26-way IDC socket at one end, and a 25-way male sub-D plug at the other end. The 26-way IDC connector is plugged on to header STL1 on the decoder board.

Pin header STL2 is inserted and soldered like STL1. STL2 provides a number of important signals in the system, and is intended for measurements and extensions. Similarly, the eight solder pins near SCART socket BU2 are intended for (optional) use of the stereo sound inputs and outputs.

As will be recalled from the circuit descriptions, the PC-VT7000 is powered from the computer. All the necessary supply voltages are carried via the I^{2}C card and the cable between this and the decoder.

I^{2}C bus controller card

The lay-out of this double-sided, through-plated board is shown in Fig. 9. The board has pre-tinned contacts for insertion into a PC extension slot.

First, fit the ten wire links, A0 through A9. Leave a little room between these wires and the PCB surface, so that they can be cut later to set the card address. Next, fit the resistors, the capacitors (four electrolytic, and six ceramic types), and then the six ICs.

Align the angled terminals of the 25-way female sub-D connector with the relevant holes in the PCB, and solder them carefully at the track side, taking care to avoid shorting adjacent pins by applying too much solder tin. Next, fit the support bracket supplied with the kit. Place it over the D-connector, and secure it with two M3 screws and nuts. In the PC, remove the rear panel bracket at the location of the slot that you intend to use for the I^{2}C card.

Before fitting the card into the PC, set its address by opening wire jumpers BR0–BR9 as required. Details on the hardware address selection are provided by the READ.ME file on the diskette supplied with the kit. Ready-assembled and aligned I^{2}C boards supplied

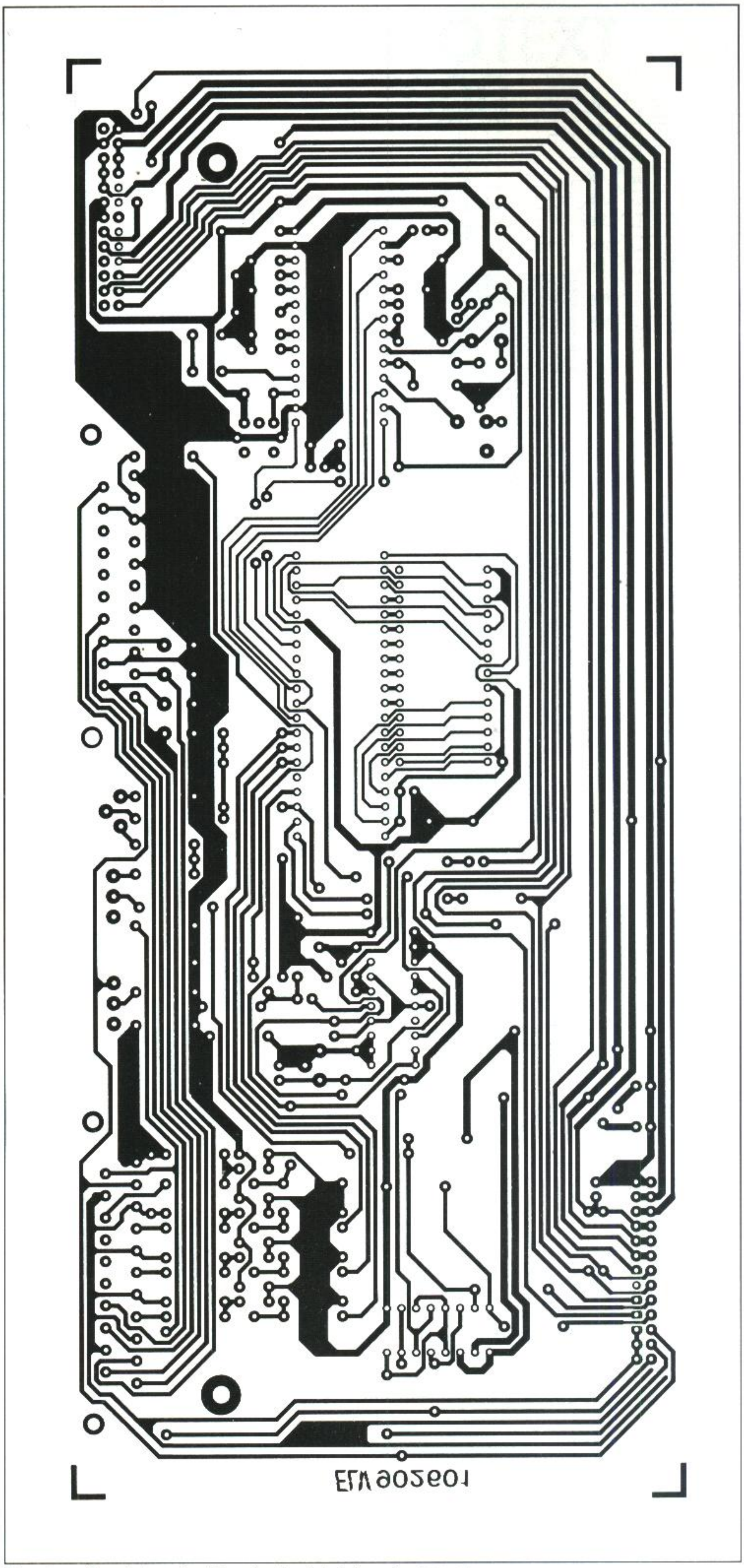


Fig. 8a. Track layout of the single-sided decoder board.

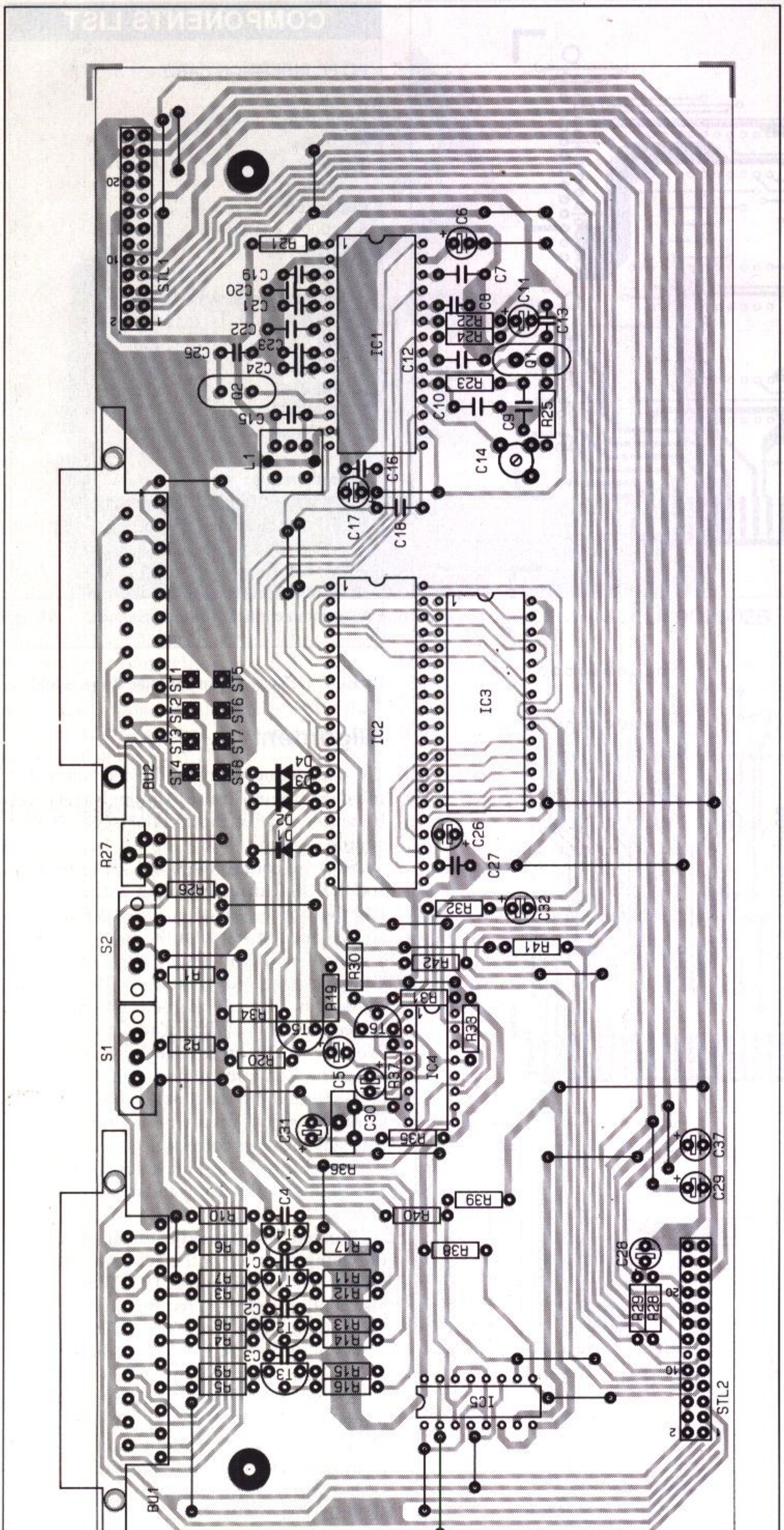


Fig. 8b. Component mounting plan of the decoder board.

COMPONENTS LIST

DECODER BOARD

Resistors:

6	75Ω	R1–R6
2	100Ω	R41;R42
6	220Ω	R7–R10;R34;R37
1	270Ω	R26
2	330Ω	R25;R32
1	470Ω	R30
3	820Ω	R12;R14;R16
3	1kΩ	R24;R28;R29
1	1kΩ2	R21
4	1kΩ5	R11;R13;R15;R35
2	2kΩ2	R17;R31
1	3kΩ3	R23
1	4kΩ7	R20
2	10kΩ	R38;R40
1	12kΩ	R19
1	68kΩ	R22
2	100kΩ	R33;R39
1	100Ω preset V	R36
1	100Ω preset V	R27

Capacitors:

2	15pF	C19;C25
1	18pF	C13
1	47pF	C15
1	100pF	C24
1	220pF	C8
1	270pF	C23
1	470pF	C21
1	1nF	C20
1	10nF	C18
1	22nF	C22
2	47nF	C10;C12
1	68nF	C7
6	100nF ceramic	C1–C4;C16;C27
1	220nF	C9
1	1µF 16V	C11
1	2µF2 16V	C6
5	10µF 25V	C17;C26;C30–C32
1	22µF 16V	C5
3	100µF 16V	C28;C29;C37
1	40pF trimmer	C14

Semiconductors:

1	SDA5231	IC1
1	SDA5243	IC2
1	6264 RAM	IC3
1	CD4052	IC5
1	CD4053	IC4
6	BC548	T1–T6
4	1N4148	D1–D4

Miscellaneous:

1	6 MHz quartz crystal	Q1
1	13.875 MHz quartz crystal	Q2
1	10µH inductor	L1
2	SPDT slide switch	S1;S2
2	PCB-mount SCART socket	BU1;BU2
2	26-way pin headers	
14	solder pin, 1.3 mm dia.	
1	printed-circuit board	
1	enclosure type 7000	
70 cm silver-plated wire		
1	Flatcable, length 2 m, 25-way, ready-assembled with D-connector and IDC connector	

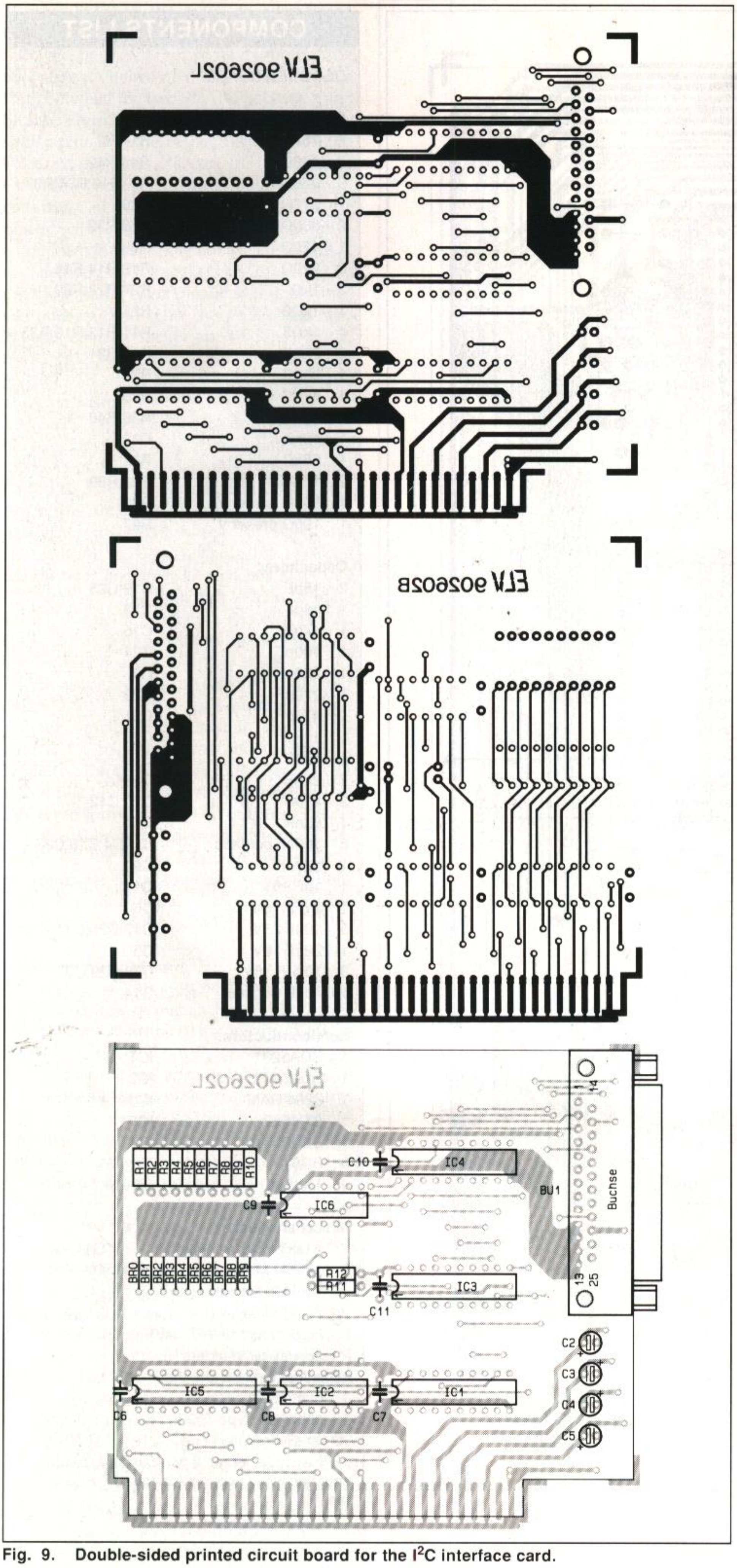


Fig. 9. Double-sided printed circuit board for the I^{2}C interface card.

COMPONENTS LIST

I^{2}C PC INSERTION CARD

Resistors:		
2	1kΩ	R11;R12
10	10kΩ	R1–R10
Capacitors:		
6	100nF ceramic	C6–C11
4	10µF 16V	C2–C5
Semiconductors:		
1	74LS32	IC2
1	74LS136	IC6
1	74LS244	IC4
1	74LS245	IC1
1	74LS374	IC3
1	74LS688	IC5
Miscellaneous:		
1	25-way PCB-mount D-connector BU1	
1	PC card support bracket	
2	M3×6 mm screw	
2	M3 nut	
12 cm	silver-pated wire	
1	printed-circuit board	

by ELV are set to operate at address 300H.

Alignment

The decoder needs to be adjusted before it is fitted into the enclosure. Although a relatively complex circuit, the decoder is remarkably simple to align.

Start by connecting a TV set to SCART socket BU1. During the alignment, the TV set has the double function of a CVBS signal source, and a display for the Teletext pages.

Run the program on the PC, and call up a Teletext page, say, number 100. If the decoder is correctly aligned, the page will appear on the TV. Without adjustment, however, the TV picture is probably unsteady — the picture may move horizontally, or may be corrupted by diagonally moving lines. Carefully adjust trimmer C14 until the picture synchronizes correctly.

Inductor L1 in the data-clock filter comes pre-aligned with the kit, and must not be adjusted.

Proceed with the adjustment by setting the clamping level for the Teletext subtitling with the aid of preset R36. The following adjustment procedure is required only if Teletext subtitles are to be recorded on a VCR.

Connect a VCR to SCART socket BU2, and switch it to stand-by. The video signal fed to pin 19 of BU2 is taken from pin 20 and routed to pin 19 of BU1 (TV). To prevent the TV switching to RGB input mode, temporarily connect pin 16 of BU1 to ground.

Use the control software to select the superimpose mode. Next, adjust preset R36 until the characters appear clearly in the picture, i.e., with the best possible contrast ratio, and without distortion or excessive brightness. Although this adjustment is perfectly feasible simply by looking at the TV picture, it may also be carried out with the aid of an

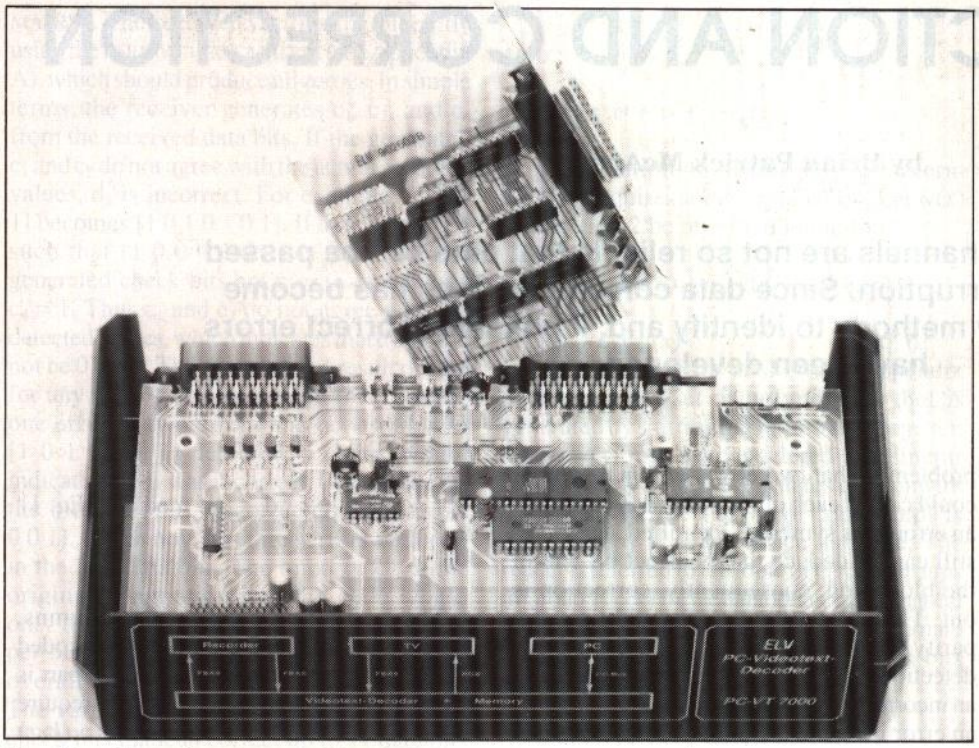


Fig. 10. Opened prototype of the decoder, and the associated I^{2}C card.

A complete kit of parts for the Videotext decoder is available from the designers' exclusive worldwide distributors:

ELV France
B.P. 40
F-57480 Sierck-les-Bains
FRANCE

Telephone: +33 82837213
Facsimile: +33 82838180

oscilloscope. Connect the scope to the collector of T5, and adjust R36 until the black level of the superimpose signal matches that of the background video signal. This completes the adjustment of the PC-VT7000.

Final assembly

If this has not already been done, mount the rear panel to the decoder board, so that the SCART sockets and the plastic pins of the two slide switches protrude from it. The flatcable is inserted via the slot provided at the right-hand side of the rear panel.

Fit the completed and adjusted decoder PCB into the lower half of the enclosure supplied with the kit. The ventilation slots are at the front side. Remove the four square blocks from the underside of the bottom half of the enclosure, and insert four M4×70 mm screws into the holes. At the inside of the enclosure, place two 1.5×10 mm dia. plastic washers over each screw at the front side, and one washer over each screw at the rear side.

The decoder board is fitted by passing the two screws at the rear of the enclosure through the two holes in the PCB. At the same time, fit the rear panel. Next, place the four 60-mm-long plastic PCB spacers over the screw ends, and mount the front panel.

Carefully lift the bottom enclosure half, and place two pencils or a folded handkerchief underneath it, so that the heads of the four screws rest on the desk while the enclosure half is a little higher. Carefully place the top enclosure half on to the lower half (the ventilation slots are at the rear side), and temporarily insert four screws into the holes. The ends of these screws partly enter the PCB spacers, and serve to keep these aligned (centred) with the holes in the top half of the enclosure. Now carefully lower the top half. Draw one corner of the enclosure over the edge of the desk, capture the screw at the underside, and push it upwards so that it ejects the screw at the top. Fit an M4 nut, and secure the screw from the underside. The other three screws are fitted likewise. Finally, push-fit the four blocks at the underside, the rubber feet, and the four square covers at the top side of the enclosure. ■

References:

1. "Video line selector", *Elektor Electronics* April 1990.
2. "Inter-IC communications", *Elektor Electronics* September 1990.

Appendix: SAA5243 ECCT register functions

D7	D6	D5	D4	D3	D2	D1	D0		
TA	$\overline{7+P}$/ 8 BIT	ACQ. $\overline{ON}$/OFF	GHOST ROW ENABLE	$\overline{DEW}$/ FULL FIELD	TCS ON	T1	T0	R1	Mode
–	BANK SELECT A2	ACQ. CCT A1	ACQ. CCT A0	TB	START COLUMN SC2	START COLUMN SC1	START COLUMN SC0	R2	Page request address
–	–	–	PRD4	PRD3	PRD2	PRD1	PRD0	R3	Page request data
–	–	–	–	–	A2	A1	A0	R4	Display chapter
BKGND OUT	BKGND IN	COR OUT	COR IN	TEXT OUT	TEXT IN	PON OUT	PON IN	R5	Display control (normal)
BKGND OUT	BKGND IN	COR OUT	COR IN	TEXT OUT	TEXT IN	PON OUT	PON IN	R6	Display control (newsflash/subtitle)
STATUS ROW $\overline{BTM}$/TOP	CURSOR ON	$\overline{CONCEAL}$/ REVEAL	$\overline{TOP}$/ BOTTOM	$\overline{SINGLE}$/ DOUBLE HEIGHT	BOX ON 24	BOX ON 1-23	BOX ON 0	R7	Display mode
–	–	–	–	CLEAR MEM.	A2	A1	A0	R8	Active chapter
–	–	–	R4	R3	R2	R1	R0	R9	Active row
–	–	C5	C4	C3	C2	C1	C0	R10	Active column
D7 (R/W)	D6 (R/W)	D5 (R/W)	D4 (R/W)	D3 (R/W)	D2 (R/W)	D1 (R/W)	D0 (R/W)	R11	Active data

– bit does not exist

R1 Mode	
T0, T1	Interlace/non-interlace 312/313 line control.
TCS ON	Text composite sync or direct sync select.
$\overline{DEW}$/FULL FIELD	Field flyback or full channel mode.
$\overline{7+P}$/8-BIT	7 bits with parity checking or 8-bit mode.
TA, TB	Test bits; 0 for normal operation.
R2 page request address	
START COLUMN	Start column for page request.
ACQ CCT	Selects one of 4 acquisition circuits.
BANK SELECT	Selects bank of 4 pages addressed for acquisition.
R4 Display chapter	Determines which of 8 pages is displayed.
R5, R6 Display control	For normal and newsflash/ subtitle.
PON	Picture on.
TEXT	Text on.
COR	Contrast reduction on.
BKGND	Background colour on.
R7 Display mode	
BOX ON 0 (1–23, 24	Boxing function allowed on row 0 (row 1–23, 24).
STATUS ROW $\overline{BTM}$/TOP	Row 25 displayed above or below main text.
R8 to R11	Active chapter, row, column and data information written to or read from page memory via the I^{2}C bus.

LOGIC ANALYSER PART 1

by K. Nischalke and H.J. Schulz

A logic analyser is primarily intended for carrying out measurements in fairly complex circuits, where an oscilloscope begins to show its deficiencies. Unfortunately, it remains a pretty expensive instrument and it is for that reason that a design is offered here that can be built by most enthusiasts at a reasonable cost.

STRICTLY speaking, it is not quite right to compare an oscilloscope with a logic analyser, because basically an oscilloscope displays voltage or current as a function of time, whereas a logic analyser shows data as a function of time. Both instruments may, therefore, be required in a number of applications. For instance, suppose that a logic analyser has indicated that the data of a certain signal are not quite right. It is then possible to check with a scope whether that is caused by the voltage behaviour of the signal (for example, a logic 1 may have become stuck at 1.5 V).

Furthermore, a logic analyser has more input channels than an oscilloscope. That is a very definite advantage when microprocessors are being tested. Even an old-fashioned 8-bit device with 16 address lines requires more than the maximum four channels available on an oscilloscope. Looking at the discrete signals associated with these devices does not mean much, but their interrelation is of great importance.

Like an oscilloscope, a logic analyser needs to be triggered, but the trigger signal must be specially adapted for working with digital information. Instead of being triggered at a certain voltage level, the logic analyser is triggered by a given pattern of bits present at the inputs.

A computer and associated software have become inseparable parts of the modern logic analyser, and in proprietary instruments they form an integral part of the construction. The present instrument is intended for operation with an external computer, either an IBM PC or compatible or an Atari.

Block diagrams

The general design of the analyser is shown in Fig. 1. The interface in the computer is connected to the busboard of the analyser. The analyser proper consists of a control board and up to four RAM cards.

The control of the analyser is an independent circuit whose paRAMeters are determined by the computer. That independence is imperative, because few computers can handle clock frequencies of 25 MHz or 100 MHz in real time. The prime tasks of the circuit are the control of the triggering and the addressing of the memory. The length of the memory is always 2 K, while its width lies between 16 bits (1 RAM card) and 64 bits (4 RAM cards).

Data are input into the RAM card(s) via a probe. The probe(s) does not consist of much more than a number of buffers, which are, however, indispensable for the correct transport of digital signals, where fundamental frequencies may be as high as 25 MHz. At the same time, it keeps the load on the circuit under test to a minimum. Furthermore, it provides a much neater connection between the analyser and the circuit under test.

The connection between the probe(s) and the RAM card(s) consists of a one-metre cable, while those between the probe(s) and the circuit under test are made by short, discrete cables. This arrangement obviates the frequently encountered spaghetti-like mess of cables. When longer connections are required, two or more probes may be connected in series.

The block diagram of the analyser, insofar as the RAM cards and the control board are concerned, is shown in Fig. 2.

When data are written via the probe(s), clock frequencies of 100 MHz, 25 MHz, 1 MHz or an external clock may be chosen. The somewhat different mode of operation at 100 MHz will be discussed later. The clock frequency determines how often the data at the inputs of the probe are clocked (sampled) in the latch at the input. Note that only in the 100 MHz mode the input latch/shift register operates as a shift register. Once the input data are sampled, they are written into RAM with the aid

SPECIFICATION

Memory	2048 samples per channel 8192 samples per channel (100 MHz mode)
No. of channels per RAM card	16 4 (100 MHz mode)
Number of RAM cards	up to 4
Clock	100 MHz; 25 MHz; 1 MHz internal) external up to 25 MHz
Clock qualifier	2 (not in 100 MHz mode); may be set high, low or disabled
Type of trigger	centre
Trigger pattern	set per channel: 0, 1 or X
Duration of trigger pattern	20 ns–5.12 µs; 'arm' triggering: 20 ns–2.56 µs
Trigger window	40 ns–81.92 µs

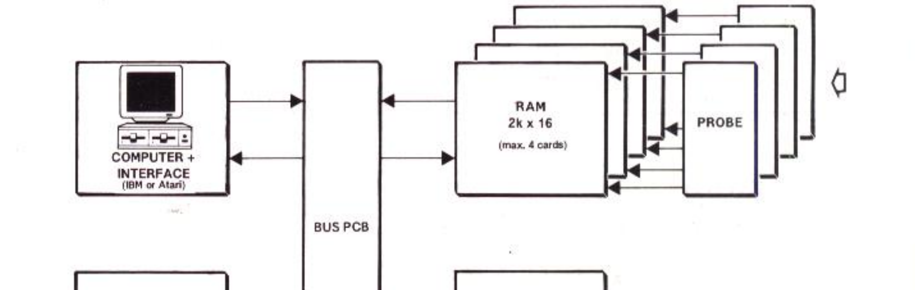


Fig. 1. General design of the logic analyser.

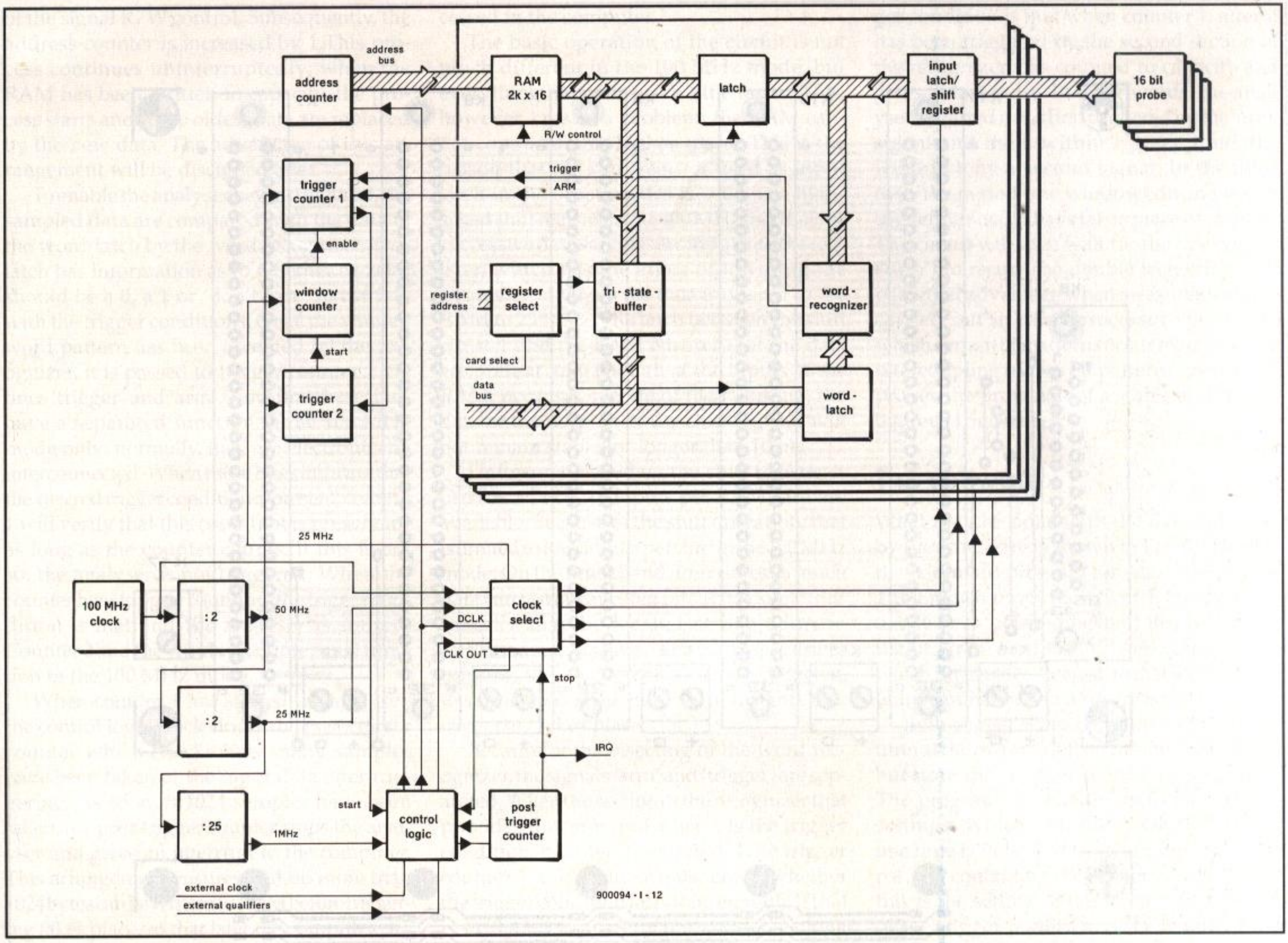


Fig. 2. Block diagram of the RAM cards and the control board of the logic analyser.

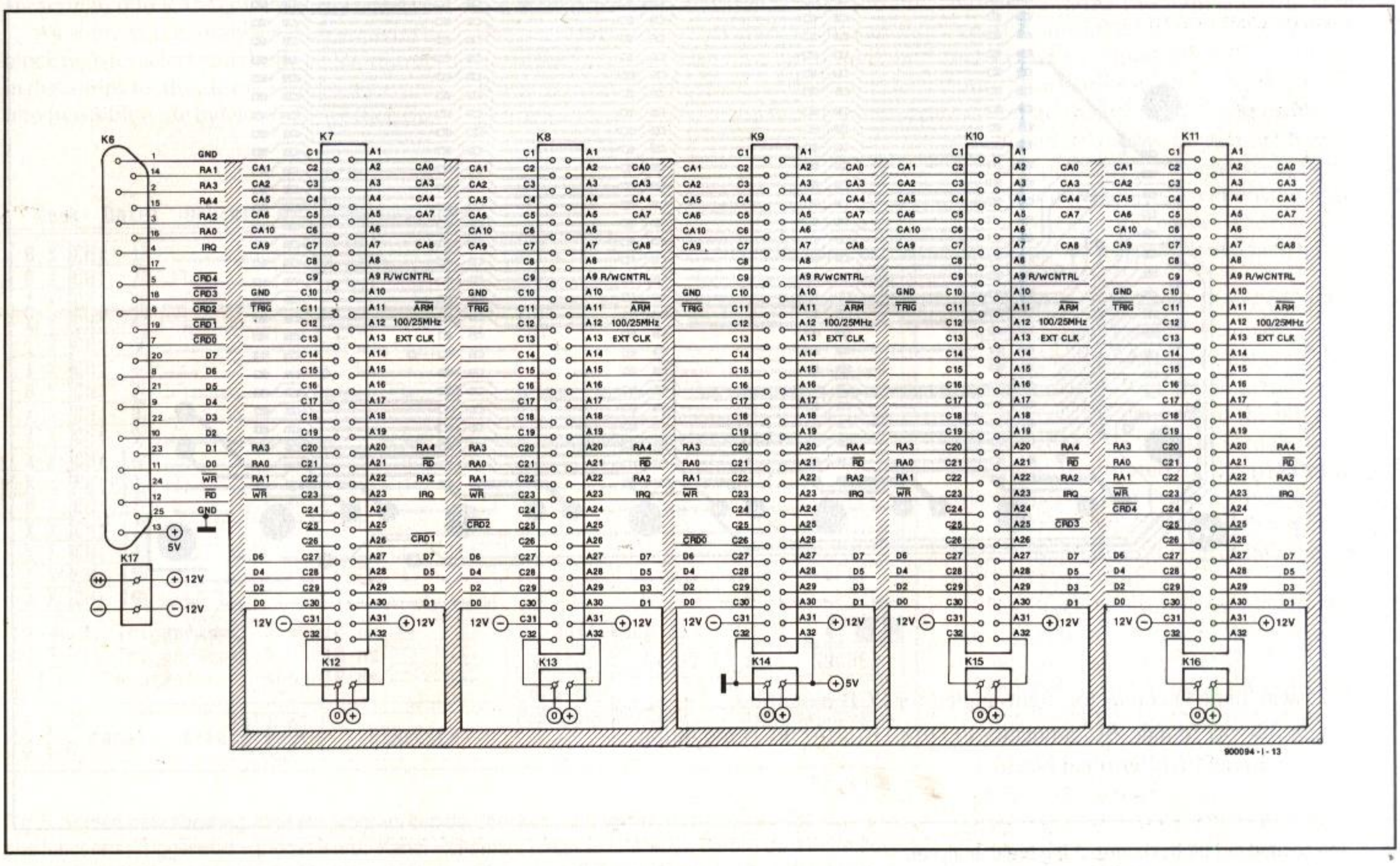
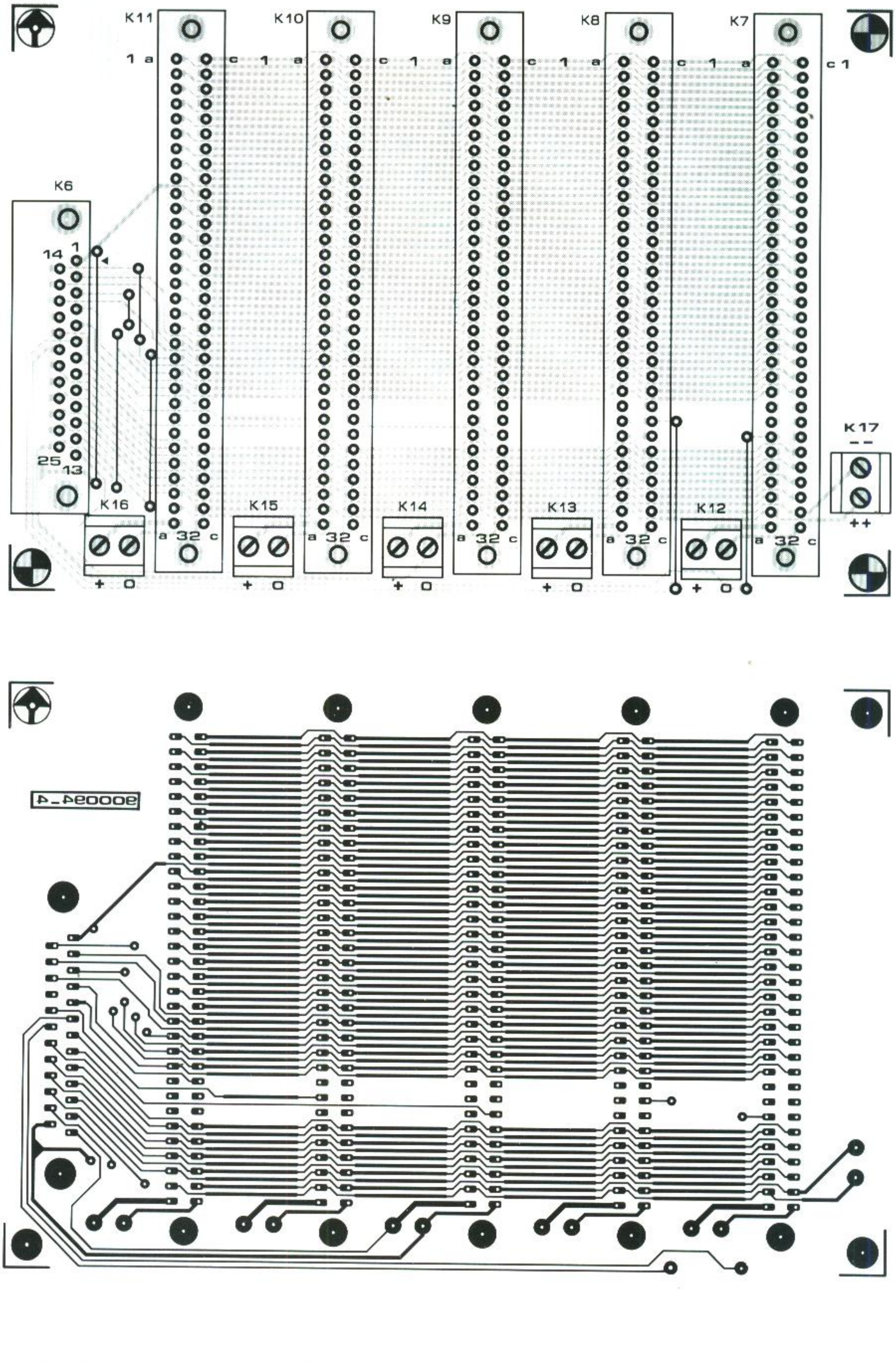


Fig. 3. The busboard is hardly more than an assembly of wiring and connectors.



K_6 = 25-way male D connector, right-angled for PCB mounting.
K_7–K_{11} = 32-way female AC connector to DIN 41612 for PCB mounting.
K_{12}–K_{17} = 2-way PCB terminal block.

Fig. 4. The printed circuit busboard of the logic analyser.

of the signal R/W control. Subsequently, the address counter is increased by 1.This process continues uninterruptedly; when the RAM has been written to capacity, the process starts anew: the oldest data are replaced by the new data. The advantage of this arrangement will be discussed later.

To enable the analyser being triggered, the sampled data are compared with the data in the word-latch by the word recognizer. This latch has information as to whether each bit should be a 0, a 1 or 'don't care' to comply with the trigger conditions. Once the wanted word pattern has been accepted by the recognizer, it is passed to trigger counter 1 via lines 'trigger' and 'arm'. Note that these lines have a separated function in the 100 MHz mode only: normally, they are electronically interconnected. When these lines indicate that the desired trigger condition is present, counter 1 will verify that this condition is present for as long as the counter counts. If this is not so, the analyser is not triggered. When the counter has stopped counting, the trigger condition is met and the analyser is started. Counter 2 and the window counter only function in the 100 MHz mode.

When counter 1 has started the analyser, the control logic block clocks the post-trigger counter which checks how many samples have been taken of the input data after triggering. As soon as 1024 samples have been taken, the post-trigger counter stops the analyser and gives an interrupt to the computer. This arrangement ensures that no more than 1024 bytes can be written to RAM before triggering takes place, so that later the computer can analyse data before and after triggering. This is the reason that before triggering the data are written into RAM continuously.

As soon as the analyser is disabled, the block register select and the three-state buffer in the computer divide the 16-bit wide words into two 8-bit wide bytes which are then processed in the computer.

The basic operation of the circuit is not much different in the 100 MHz mode, but everything happens just a little faster. This, however, creates a problem: the RAM cannot cope with this higher speed. This is the reason that the shift register used as input latch in the other modes is used as a high-speed shift register in the 100 MHz mode. Four successive data samples are input into the register, which has the effect of lowering the frequency at which the data is offered to the RAM to 25 MHz. The latch between the shift register and the RAM ensures that the data are present long enough at the inputs of the RAM, because, in spite of their shifting, the data at the outputs of the shift register will not remain stable for longer than 10 ns.

Unfortunately, when the shift register is used as a latch, 16 inputs per RAM card are available. Because of the shift operation, that is limited to four inputs per card in the 100 MHz mode. On the other hand, four times as much data can now be written into RAM since four samples may be located at one address. Furthermore, because of the lower number of inputs, the word recognizer may be bisected, which increases the number of instants that triggering takes place.

Because of the bisecting of the word recognizer, the signals 'arm' and 'trigger' are separated. When the section of the recognizer that provides the 'arm' pulse accepts the trigger condition, counter 2 is started. Like trigger counter 1, this counter must check whether the trigger condition lasts long enough. If that is so, the window counter is started. As long as this counter runs, three situations may occur: (a) the 'arm' trigger word appears again; (b) the second trigger condition is met; and (c) the window counter has counted to capacity. In the first case, trigger counter 2 and the window counter are reset, so that the triggering process starts anew. The second trigger condition is met when counter 1, after it has been triggered by the second section of the recognizer, has counted to capacity and gives a start pulse. In other words, the analyser is started if it is first triggered by the 'arm' signal and then, within a given time (the window), by a second signal. In the third case, the period (the window) during which triggering should have taken place has lapsed. The circuit will then wait for the first trigger ('arm') to recur. The double triggering is of particular advantage when measurements are carried out in microprocessor systems in which certain bit patterns occur regularly. The intercoupling of two bit patterns greatly improves the provision of a stable and unambiguous triggering.

The software

What may be done with the data collected by the hardware is shown in Fig. 5, which is the (German) program for Atari computers. The English translation, which has unfortunately been delayed beyond this issue, and the program for IBM PCs and compatibles, which is virtually identical to that for the Atari, will be published in a future issue.

Both trigger words may be set in the column at the extreme left on the monitor screen, but note that trigger word 2 is not active. The program can handle up to four trigger settings. Which of these is selected at any one time is indicated by the figure under the column containing the trigger words. Beside this is the setting of the minimum duration of the trigger pattern and the length of the trigger window.

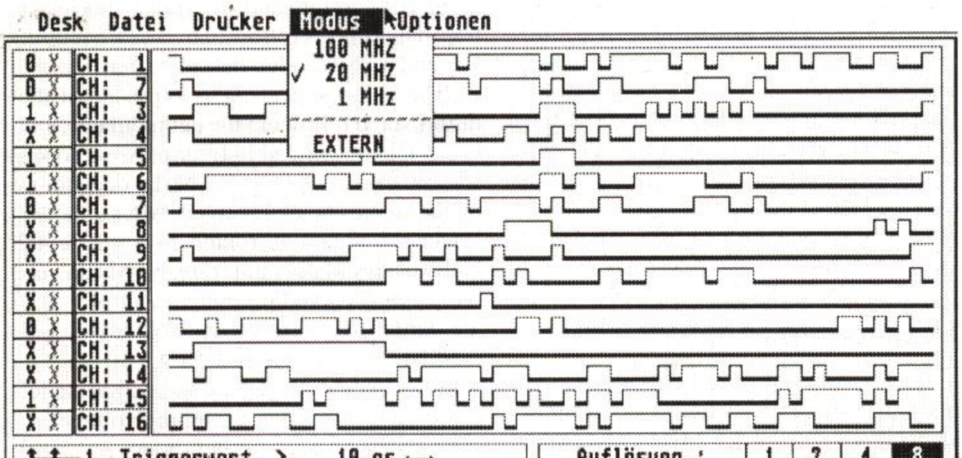
The column beside that for the trigger words contains the names of the signals. Apart from the standard indication "channel so and so", these signals may be given more informative names. Also, it is possible to arrange them in a different order on the screen and to display them more than once. For example, associated signals may be placed side by side on the screen without the necessity of changing over test probes. The column may also be used to indicate from which test probe the signal comes.

The larger part of the screen is, of course, reserved for the display of signals, and it can handle up to 16 of these simultaneously. Other channels may be brought on to the screen by scrolling via the window at the bottom left. The control for shifting the signals to the left and right is situated at the bottom right. Above that is an indication where the (so far invisible) cursor is situated with respect to the triggering point, and which data (hexadecimal) are portrayed.

The block 'solution' enables the determining of how many samples are required on the screen simultaneously.

The pull-down menus provide the following functions: (a) information about the program (desk); (b) storing on disk of the data to be analysed (datei); (c) printing of the data (printer); (d) selecting the clock frequency (mode); and (e) various functions (options).

(to be continued)



0 X CH: 1
0 X CH: 7
1 X CH: 3
X X CH: 4
1 X CH: 5
1 X CH: 6
0 X CH: 7
X X CH: 8
X X CH: 9
X X CH: 10
X X CH: 11
0 X CH: 12
X X CH: 13
X X CH: 14
1 X CH: 15
X X CH: 16

1. Triggerwort > 10 ns
2. Triggerwort > 10 ns
Distanz < 10 ns
1

Auflösung : 1 2 4 8
Status : 64 69 76 00
Abstand : 0 ns

Kanal: 1-16
⇧ 4 ⇧ ⇩ 4 ⇩
⇧ 1 ⇧ ⇩ 1 ⇩

Fig. 5. Screen view showing what the program can do. Drucker = printer; Modus = mode; Optionen = various functions; Triggerwort = trigger word; Kanal = channel; Einlesen = to write; Datei = data in; Auflösing = resolution; Suchen = to search; Abstand = distance. (Proper English translation will be published soon).

23/24-CM FM ATV RECEIVER

This review deals with the recently introduced 23/24 cm FM Amateur Television (ATV) receiver from EHC (Valves) Ltd. At the time of writing, it is the only complete system that can be purchased in a 'ready-to-go' state. That is, all you have to is connect the DC shack supply, plug in a 24-cm aerial, a video monitor and a loudspeaker or headphones, tune in a signal, and away you go.

by Mike Wooding, G6IQM

ALTERNATIVELY, the IF board and the down converter sub-sections can be purchased separately for inclusion in your own system, or for building your own version of the receiver. However, this option will require a number of external components such as a meter, controls, sockets, etc., to be added.

Description

The receiver is housed in a grey coloured moulded impact-proof PVA plastic box measuring 23.5×18.25×9.5 cm including the feet, with the control knobs on the front panel, and the sockets on the rear panel making the unit 21.5 cm deep in total. The case also features adjustable front feet, allowing the unit to be tilted upwards to facilitate reading the meter.

The front panel controls comprise, from the left to the right: an on/off switch, a TUNING control, a VID. GAIN control and a VOLUME control. Also featured on the front panel is a red LED for ON/OFF indication, and the tuning-frequency meter. The front panel bezel is sign-written with the various control and switch functions.

The rear panel connections are as follows: a 3.5-mm jack socket for the loudspeaker/headphone connection, a phono socket for video output, a BNC socket for aerial connection, and a DC socket for the power supply input. Also mounted on the rear panel is a 20-mm fuse holder.

Internally, the receiver has a die-cast box and two printed circuit boards mounted onto the housing chassis. All earth connections are bonded to this chassis for screening security.

The die-cast box, measuring 120×95×35 mm, houses the down converter with power connections made via feed-through capacitors, and the IF output by means of a coaxial cable. The aerial input socket mounted through the rear panel of the housing is a bulkhead coaxial cable connection type, connected to the down converter by a short length of coaxial cable.

The two PCBs are the wide-band IF amplifier and the IF receiver assemblies. Both boards are constructed from double-sided copper-clad material, and are custom-built for each unit. All interconnecting wires and cable are neatly formed and retained with cable ties.

The TUNING control is a precision 10-turn potentiometer, which facilitates fine-tuning of the band with only one control. The VID. GAIN and VOLUME controls are ordinary 270-degree potentiometers. The tuning frequency meter is a standard 90-degree movement type with the scale calibrated with a logarithmic frequency read-out scaled in 10-MHz steps, with main divisions at 1240, 1260, 1300 and 1320 MHz.

Circuit outline

The number of controls has been kept to a minimum, i.e., tuning, video gain, volume and changeover from 5.5 MHz to 6.0 MHz sound subcarrier. This latter switch also changes the polarity of the video output from negative on 6 MHz (UK and most of Europe) to positive on 5.5 MHz (France).

The first down converter section (housed in a die-cast box) uses parallel-tuned circuits to give a good rejection of broadcast TV signals, radar and other in-band signals. All the SHF and UHF tuned circuits are etched on to the printed circuit board, including the first local oscillator, which reduces frequency drift to a very low level after only a few minutes warm-up. The output from the first down converter is 168 MHz, which is routed to a second down converter. This is external to the die-cast box, and comprises a dual-IC tuned circuit that converts the first IF of 168 MHz to a second IF at 39 MHz for input to the demodulator.

The 39-MHz signal is amplified by another dual-IC circuit and then buffered by an emitter follower stage, which matches the signal to the phase-locked loop demodulator. After demodulation, the video signal passes through an emitter follower buffer stage, a standard (non-switchable) CCIR de-emphasis circuit, and an audio subcarrier trap. The latter is followed by a video amplifier and a 75-Ω video output buffer.

The audio subcarrier is extracted at the buffer stage following the demodulator, and is fed into a switchable filter network to select either 5.5 MHz or 6.0 MHz, depending on the signals being received. After the filter stage, the signal is routed to a single IC audio demodulator, and from there to an output stage. The audio volume control varies the output from 0 to about 1 W into an 8-Ω load, or 2 W into a 4-Ω load.

The receive frequency indication meter is driven by sampling the tuning voltage applied to the varicap tuning diode. A potentiometer network allows the meter range to be adjusted to compensate for the different characteristics of different varicap diodes.

Bench test

In view of the fact that the receiver is supplied as a complete unit, the nature and depth of the bench test were different compared to other reviews I have conducted in the past. However, what I consider to be essential tests I was able to carry out, and the results are discussed below.

The **noise floor** of the receiver with the aerial input terminated with a 50-Ω load was averaged out at −85 dBm. Three spurious signals were found at 1116 MHz (signal level: −25 dBm), 2232 MHz (−35 dBm) and 3348 MHz (−60 dBm). However, although they were at not inconsiderable levels, there did not appear to be any detraction in the performance of the down converter.

It ought to be said here that in virtually all circuits employing these relatively simple, but stable, oscillators working at the direct conversion frequency, harmonics and spurious signals will inevitably be produced. The trick is to tailor the circuitry such that they are either suppressed, or at frequencies outside the operating range of the receiver itself. Also, care must be taken that these harmonics and spurious signals do not desensitize the front end of the converter, especially when this is based on GaAs-FET devices and strip-line techniques.

The **sensitivity** of the receiver was checked by injecting signals from the Racal Dana 9087 Signal Generator, and tracking the output of the second IF stage with the Marconi 2383 Spectrum Analyser. With an input frequency of 1249 MHz, a signal level of −71 dBm was required to give an IF signal 3 dB up on the noise floor, or, in other words, just about the minimum signal level that could be detected as sync pulses. At the other end of the band, 1320 MHz, a slightly lower input of −72 dBm was required to give a 3-dB lift. These signal levels of around −70 dBm equate to a performance of the receiver similar to that of most of the equipment at present operated by ATV-ers.

The **maximum signal levels** before the IF stages went into limiting were −29 dBm at 1249 MHz, and −27 dBm at 1320 MHz, which indicated a good large signal handling capability (−30 dBm is equivalent to 1 mW).

Being unable to apply an external modulating signal greater than 100 kHz to the Racal signal generator, I had to inject a modulated signal into the demodulator itself at the second IF frequency from the Philips PM5646 Television Pattern Generator, and view the output on a video modulator. The resulting picture showed good definition and a linear chrominance response.

Using the same method as above, a 1-kHz audio signal was also injected from the test pattern generator, and the resultant audio signal measured on the HP 8903B Audio Analyser. The frequency recovery was found to be excellent, with the measured output signal reading 998 Hz. The distortion level at maximum output into a 8-Ω load was measured at 3.154%.

The **current consumption** of the receiver at a supply voltage of 13.5 V was measured at about 225 mA with no audio output, and 520 mA at full audio output.

Off-air tests

To conduct the off-air tests I simply compared the unit against my own Wood & Douglas/BATC home brew system, which has given me excellent results in the past. Having two 24-cm aerials of the same type and at the same height, and separate video monitors made this task easy.

First, I used the signals from GB3RT, the Coventry 24-cm FM ATV repeater, which I receive normally at P5 (broadcast quality). The pictures from the review unit were just as good as my own system, once the correct video output level was obtained with the front panel control. The tuning control, being a multi-turn type, allowed for easy tuning across the signal, enabling me to fine tune with considerable ease. The picture did have a tendency to bounce about somewhat until the correct tuning point was achieved, and the demodulator took a second or two to stabilize. This I took to indicate that owing to the strength of the received signal the IF stages were in limiting, hence perhaps slightly overloading the following stages. This has to be tempered a little by explaining that the signals I receive from GB3RT, although it is some 16 km away, are very strong indeed.

Next, I turned the aerials northwards towards GB3GV, the Leicester 24-cm FM ATV repeater, which I normally receive at P3 to P4. The results from the review unit were marginally worse than my own unit as far as the strength of the received picture itself was concerned, but the actual definition of the picture itself was, to all intents and purposes, identical.

Having checked at the top of the band, it was time to see how it fared where most of us ATV-ers operate, at the bottom end, around 1249 MHz. For this purpose I enlisted (as usual) the good offices of my fellow GB3RT group members Tony G0HOV, and John G1IJT. The received pictures all exhibited good video definition and chrominance response (apart from Tony's camera, but that's another story). By moving the aerial away from each station I was able to simulate very low received signals, and only at very low levels was the review unit found to be less sensitive than my own, but this was at picture strengths barely reportable as P1 (corresponding to just discernible prominent features, locked syncs).

The final test was to monitor my own transmissions. Again with the aid of Tony as the receiving station (for licence complicity purposes, you understand) I transmitted at an output level of 2.5 W from my basic transmitter, and then at 60 W with my linear amplifier connected in. The results from the review unit at both power levels were the same, and quite a bit worse than with my own system. The resultant video output was quite unstable at times, with obvious sync crushing taking place in the overloaded demodulator. As with my own system, there was also a multitude of points along the tuning scale where my transmissions could be found.

TEST EQUIPMENT USED FOR THIS REVIEW

- **Marconi 2383 Spectrum Analyser and Tracking Generator**
- **Racal Dana 9087 Signal Generator**
- **Philips PM5646 Television Pattern Generator**
- **Hewlett Packard 8903B Audio Analyser**
- **Philips PM3226 Digitising Oscilloscope**
- **Fluke 8050A Digital Multimeter**
- **Racal Dana 9232 Bench Power Supply**

Perhaps this test is a little unfair in expecting the unit to perform without fault, but is must be remembered that this is the *modus operandi* in many an ATV shack. The ability to monitor your own transmitted signal is a must, and most of us admit to doing it by using our own receivers, rather than 'sniffing' the output RF and demodulating that back to video.

Conclusions

Overall I quite liked the unit. As it stands, it is at present the only complete receiver available fully built, aligned and ready to go. The presentation is tidy, and the operation simplicity itself.

I do not really like the external video output level control. I feel that the output should be internally preset at 1 V at 75 Ω, which is the international standard. This would overcome the tendency of many people to incorrectly drive their video monitors and cause poorer pictures to be displayed.

I also feel that the CCIR de-emphasis network should be made switchable. There are many stations who do not use pre-emphasis in their transmissions, or at best, incorrect levels of emphasis. To be able to select the de-emphasis network allows for the correct reception of such signals. Otherwise, poorly defined and blurred pictures result, which hardly helps to boost confidence in the receiver.

The FM ATV receiver is available from EHC (Valves) Ltd., 7 Pavement Square, Lower Addiscombe Road, Croydon. Telephone: (081 654) 7172, at a cost of £230.00 including VAT. Carriage is extra at £3.34.

Finally, I wish to thank Mr. Brian Aylward of EHC (Valves) Ltd., for the loan of the review unit, and for his advice and assistance. ■

THE COMPLETE PREAMPLIFIER (1)

by T. Giffard

After the several 'spartan' preamplifiers we have published over the past few years, we feel the time has come to offer our readers a more sophisticated design, one with more controls than just a selector switch and a volume control. Although the design presented has quite a few useful features, and offers very good signal handling qualities, we have been able to keep its price to a reasonable level.

LOOKING around in hi-fi dealers' and retailers' shops (and talking to the manager or owner), it appears that a majority of (prospective) preamplifier buyers are intertested in at least some of the following features. First on the list is a copying facility that is independent of the selector switch; second, the possibility of recording from one recorder to another; third, a tone control with variable cut-off frequencies; fourth, a headphone output (which means that the main amplifier can be switched off if listening only via the headphones is required); fifth (and surprising in these days of the Compact Disc), a dynamic pickup input. And, of course, the preamplifier must not degrade the signal in any way.

With these requirements, and that of low cost, in mind, we have come up with a unit that meets them handsomely and still offers the possibility of using ICs of different price classes.

The design lends itself to fairly easy construction. The source and record selector switches are housed on a busboard, together with the input and output connectors, while the remainder of the electronics is fitted on a motherboard. The controls are connected to the circuitry via a number of terminals that are situated at the front edge of the motherboard.

Overview

The block diagram of the preamplifier—see Fig. 1, shows that there are six inputs, of which one is the dynamic pick-up input. If that input is not needed, it may be transformed to a normal line input by a simple wire bridge. The CD input has provision for adding a board (for instance, a digital-to-analogue converter) at a later date.

The input signals are fed to two rotary switches: one for selecting the input mode (i.e., the signal that will be processed, and the other for selecting a signal to be recorded. In that way, it is possible to record from a CD and continue listening to the tuner. Standard rotary switches are used since these are much cheaper than the 12 relays that would otherwise be necessary. The switches are located immediately adjacent to the the inputs at the rear of the enclosure, so that long signal paths are avoided.

Either switch is followed by a buffer, which in turn is followed by a stereo/mono switch which uses a summing opamp. This obviates the level differences that frequently occur in stereo/mono systems.

The tone control is a low-high design with two cut-off points at either side. The control range has been kept fairly limited to avoid overdriving of subsequent stages; it is, however, more than adequate for normal usage and offers smooth operation. For those who do not want tone control in any circumstances, it may be taken out of circuit by a simple 'tone defeat' switch.

The balance and volume control are followed by the output amplifier, which provides the only amplification of the line signals. The opamp chosen for this stage is able to drive loads of 600 Ω and higher.

A relay providing a delay at switch-on is provided at the output: this gives the unit a few seconds after the supply is switched on to stabilize during which period no signal is applied to the output.

The output amplifier also supplies part of the signal to a special headphone amplifier that consists of an opamp and discrete output stage. It provides enough power for driving low-resistance headphones.

The circuits

The blocks of diagram Fig. 1 are easily recognized in the circuits of Fig. 2. and Fig. 3. The circuit has been split into two to correspond more closely to the drawings of the associated printed-circuit boards.

Figure 2 shows the record and source switches S_1 and S_2 with associated buffers, IC_1 and IC_2, and all inputs and outputs with the exception of that for the headphones. Each input is shunted by a potential divider, for instance, R_1-R_2 for the lefthand CD channel. These dividers largely determine the input imedance of 47 kΩ. It is advisable to use the dividers only if really necessary, since they may adversely affect the cross-talk between the left- and right-hand channels, as well as between the inputs. If dividers are not used, resistors R_1, R_3, R_5, and so on should be replaced by wire bridges.

The CD input has some additional facilities. Normally, the analogue CD signal is

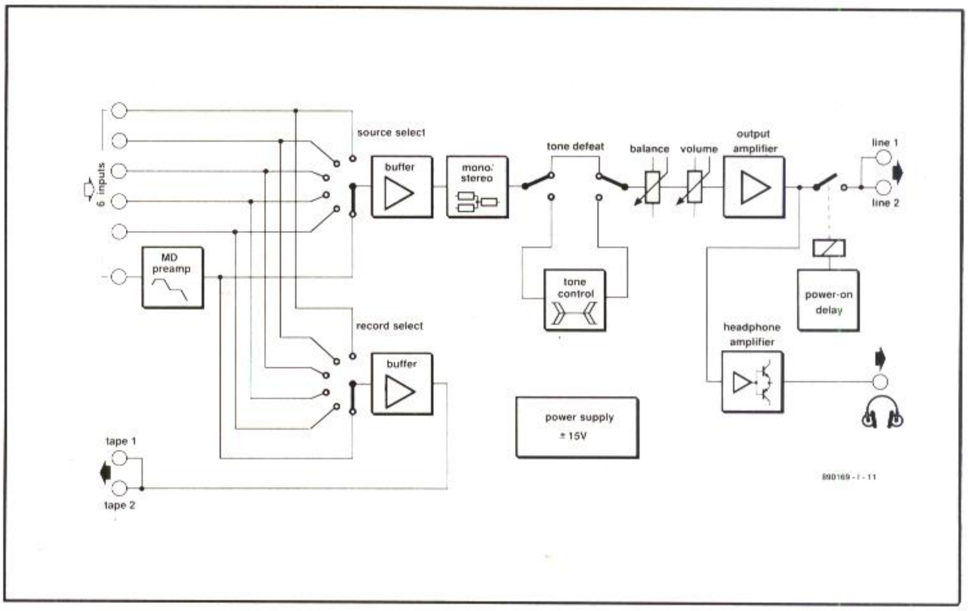


Fig. 1. Block diagram of the preamplifier.

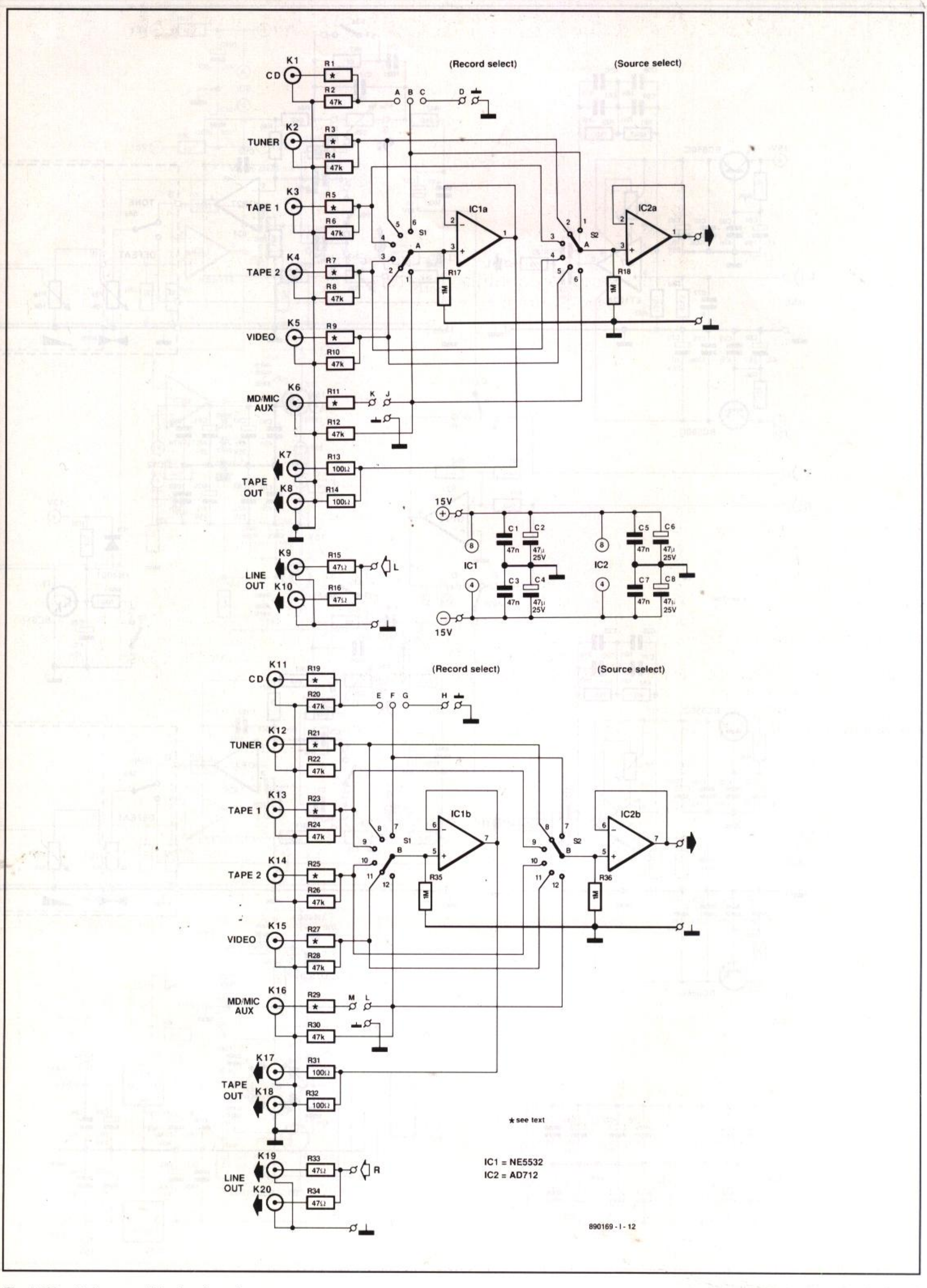


Fig. 2. Circuit diagram of the busboard

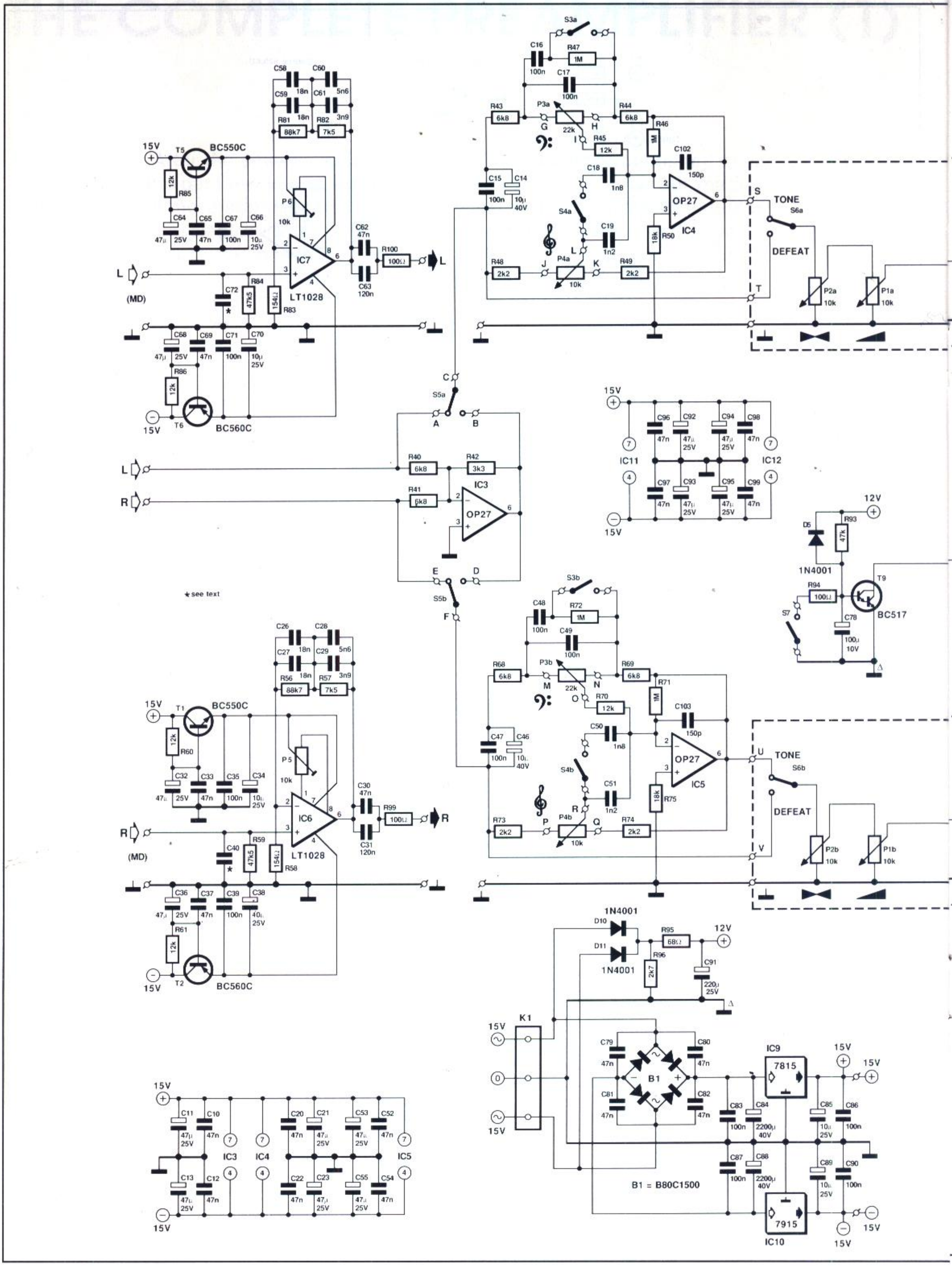
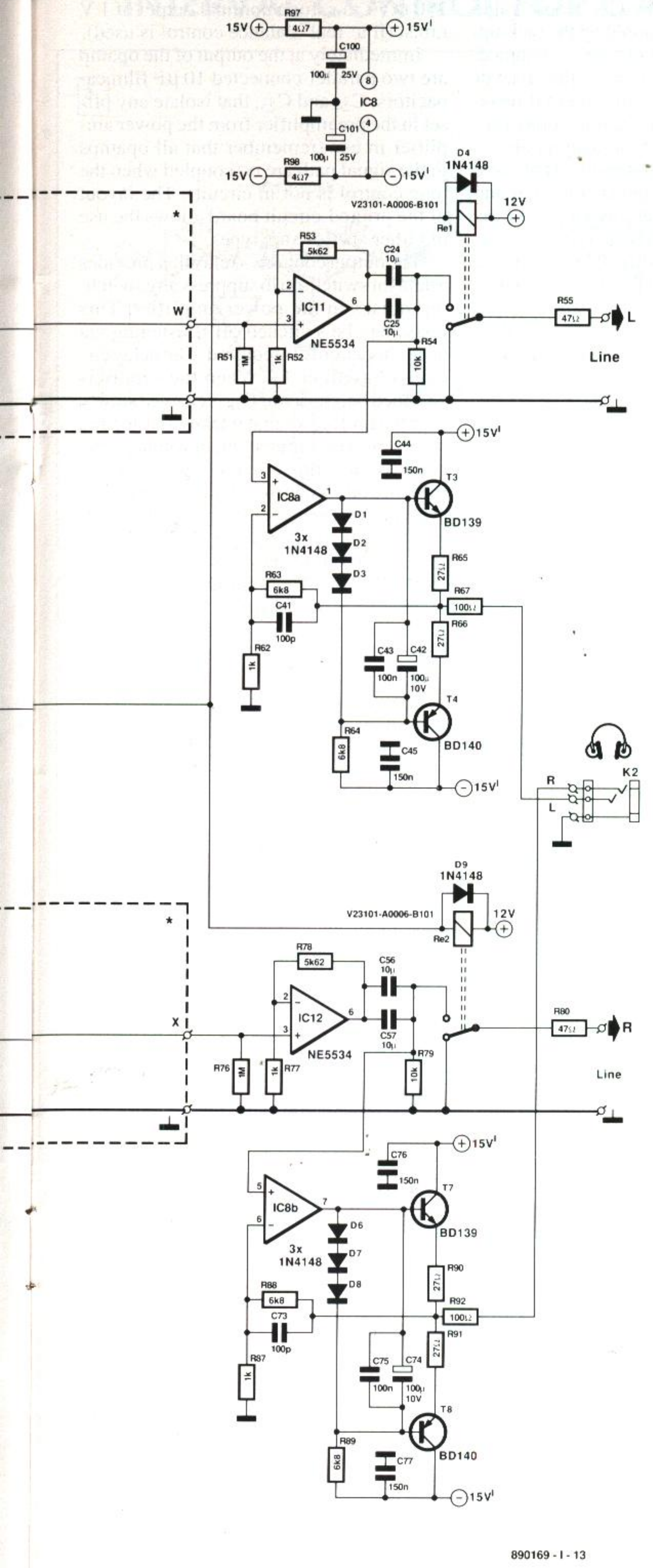


Fig. 3. Circuit diagram of the motherboard.


15V
R97
4Ω7
15V'
C100
100µ
25V
8
IC8
4
C101
100µ
25V
R98
4Ω7
15V
15V'
D4
1N4148
V23101-A0006-B101
12V
Re1
*
R53
5k62
C24
10µ
C25
10µ
W
2
3
IC11
6
NE5534
R54
R55
47Ω
L
Line
R51
1M
1k
R52
10k
15V'
C44
150n
T3
3
2
IC8a
1
D1
3x
1N4148
D2
D3
BD139
R65
27Ω
R63
6k8
R67
100Ω
C41
100p
R66
27Ω
R62
1k
C43
100n
C42
100µ
10V
T4
R64
6k8
C45
150n
BD140
15V'
R
K2
L
D9
1N4148
V23101-A0006-B101
12V
Re2
*
R78
5k62
C56
10µ
2
IC12
6
X
3
C57
10µ
NE5534
R79
R80
47Ω
R
Line
R76
1M
1k
R77
10k
15V'
C76
150n
T7
5
IC8b
7
6
D6
3x
1N4148
D7
BD139
D8
R90
27Ω
R88
6k8
R92
100Ω
C73
100p
R91
27Ω
R87
1k
C75
100n
C74
100µ
10V
T8
R89
6k8
C77
150n
BD140
15V'
890169 - I - 13

applied direct to the CD input. A wire bridge between A and B (E and F) feeds the signal to the two rotary switches. Terminals C, D and ground (G, H and ground) enable a digital input, coupled to a separate digital-to-analogue amplifier, to be provided at a later stage.

The dynamic pick-up input also has additional facilities. Normally, the associated preamplifier on the motherboard is connected between K and J (left-hand channel) and between M and L (right-hand channel). If a dynamic pick-up input is no longer required, the input may be used as microphone input, for which the preamplifier is easily adapted, or as line input, in which case the preamplifier is not used and J-K and M-L must be linked by wire bridges.

The opamps shown in the circuit diagram are not necessarily the cheapest types, but they are the best for the particular purpose as far as our prototypes showed. This aspect will be touched upon again in Part 2 under Construction. For the moment, as an example, buffer IC_1 is a Type 5532, an inexpensive yet excellent double opamp, whereas buffer IC_{2a}, a Type AD712, is a much dearer type. This type was found necessary to obviate audible switching noises when S_2 is turned. These noises result from the change in total resistance at the input of the opamp: if, for instance, the CD input is used, R_{18} is shunted by R_2. If an opamp with transistor inputs, such as the NE5532, were used, the change in resistance would cause a corresponding change in the output of the opamp. An opamp with FET inputs, such as the AD712, hardly reacts to the resistance change. If notwithstanding this a Type 5532 is used, R_{16} and R_{36} should be reduced to 220 kΩ. The input impedance will then reduce to about 39 kΩ, and this means that the ratio of the potential divider will change slightly.

Apart from the inputs, the busboard also contains all the outputs: two tape outputs with an output impedance of 100 Ω, determined by R_{13}, R_{14}, R_{31}, and R_{32}; and two line outputs, also with an output impedance of about 100 Ω, determined by R_{15}, R_{16}, R_{33}, and R_{34}, and R_{55} and R_{80} on the motherboard.

The circuit contained on the motherboard is shown in Fig. 3. Since the design of the left- and right-hand channels is identical, only the left-hand one will be referred to in what follows.

The preamplifier for the dynamic pick-up is shown somewhat away from the main circuit to emphasize its short connections to terminals J-K and L-M on the busboard. The opamp chosen for this stage is a very-low-noise Type LT1028. If that is found too expensive, the cheaper OP27 or 5534 may be used. The RIAA (IEC) correction is provided by R_{81}–R_{83} and C_{58}–C_{61}. Network R_{12}-C_{62}-C_{63} forms a high-pass filter with a cut-off frequency of 20 Hz to conform to the IEC requirement. If that frequency is thought unnecessary, the values of C_{62} and C_{63} may be increased.

The input of the preamplifier for the dynamic pick-up is entirely DC-coupled to obviate any degradation of the signal by coupling capacitors. The input capacitance is determined primarily by C_{72}, whose value depends on what is required by the pick-up. If that is not known, 47 pF may be assumed. Because of the direct coupling, the offset of the opamp depends on the internal resistance of the pick-up and P_6 is therefore provided to cancel the offset as appropriate.

The supply lines to this sensitive stage have been provided with additional filters (around T_5 and T_6) to even out any ripple on the ±15 V supply. It should be noted in this context that the use of an LT1028 is justified only in conjunction with low-impedance signal-sources (<400 Ω). If a standard high-impedance dynamic pick-up is used, an OP27 will be perfectly all right, since the total noise then consists primarily of the thermal noise emanated by the pick-up.

The mono-stereo switch, S_5, is preceded by a summing amplifier, IC_3, that provides true addition of the left- and right-hand channels without any attenuation when stereo signals are to be processed as mono ones. Passive circuits always introduce some attenuation.

Although the design of the tone-control stage, IC_4, appears conventional, it has some interesting aspects. Usually, the connections to the ends of the carbon tracks of the 'high' potentiometers are via capacitors that determine the onset of the control. If the onset frequency is to be altered, the value of both capacitors needs to be changed. In the present design, only one component per channel needs to be altered. Switches S_3 and S_4 enable two different onset frequencies to be selected. The circuit is consequently somewhat simpler as regards wiring and switching.

A 1 MΩ resistor, R_{45}, provides a feedback loop that prevents the output of the opamp from rising to unwanted heights if the wiper of the 'low' potentiometer occasionally loses contact with the carbon track. Capacitors C_{14} and C_{15} have been added to ensure that the offset of the input buffers is not amplified in the tone control stage. This parallel network of a bipolar and a film capacitor guarantees faithful signal processing even at high frequencies. The control range for various onset frequencies is shown in Fig. 4 (published in our next issue). If required, the entire tone control may be taken out of circuit by S_6.

It is recommended to use the very best quality potentiometers for the balance and volume controls, for instance, Alps types. These are not cheap, but they give consistently good performance over a long life. In some locations, a real balance control may be difficult or even impossible to obtain: the right type has half-silvered tracks so that no attenuation occurs in the centre position. If so, a standard linear stereo potentiometer may be used with a 2k2 resistors soldered between its wiper and the pole of S_{6a}: the attenuation in the centre position is then minimal.

The line amplifier, IC_{11}, is a Type NE5534, which has the advantage of being able to deliver more than 8 V r.m.s. into a 600 Ω load. Its amplification has been set at ×6.6 to give a sensitivity at the line inputs of 150 mV r.m.s. for a nominal output of 1 V r.m.s. (if a 'real' balance control is used).

Immediately at the output of the opamp are two parallel-connected 10 μF film capacitors, C_{24} and C_{25}, that isolate any offset in the preamplifier from the power amplifier in use (remember that all opamps in the signal path are DC coupled when the tone control is not in circuit). The layout of the printed-circuit board allows the use of either MKT or MKP types.

The output contains a relay that provides a delay at switch-on to suppress any switching noises in the power amplifier. This relay may be switched off if listening via headphones only is required. The delay circuit is based on T_9. When the supply is switched on, capacitor C_{78} is charged slowly via resistor R_{93}, so that it takes a little time before the base-emitter junction voltage has reached the value at which the transistor switches on. When the supply voltage is removed, C_{78} discharges fairly rapidly via D_5 and the relay is deenergized virtually instantly. The circuit has its own rectifier network, R_{95}-R_{96}-C_{91}-D_{10}-D_{11}, that makes rapid switch-off possible.

The headphone amplifier consists of a Type 5532, IC_{8a}, and a discrete complementary output stage, T_3 and T_4. Three diodes, D_1–D_3, ensure correct class-A operation. The amplifier delivers sufficient power for driving low-impedance and insensitive headphones (with the exception of electro-static ones that often require a real power amplifier). The 100 Ω resistor in the output limits the maximum current, which can rise to 100 mA with an output impedance of 8 Ω. The amplification of the stage has been set to a value at which clipping of the line output voltage of 1 V just does not occur.

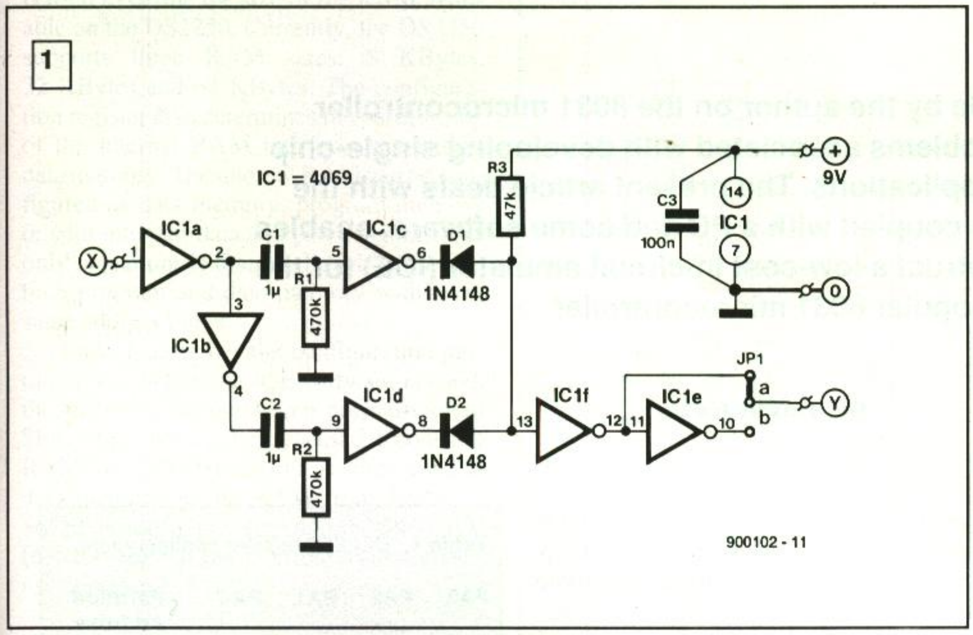
The supply of the output stages, IC_8, T_3, T_4, T_7, and T_8, has its own decoupling, provided by R_{97}, R_{98}, C_{100}, and C_{101}, to prevent any feedback to the other opamps in the amplifier when the output current is high.

The power supply of the preamplifier is simplicity itself: a bridge rectifier, B_1, decoupled by capacitors C_{79}–C_{82}, buffer capacitors C_{84} and C_{88}, and two regulator ICs, IC_9 and IC_{10}. Note that each IC in the preamplifier has additional supply decoupling in the form of an electrolytic and a ceramic capacitor.

(to be continued)

INTERVAL CONTROL FOR CAMCORDERS — AN UPDATE

1



THE original design of an interval control for camcorders (Ref. 1) works with camcorders that are switched on when the RECORD switch is pressed, and off when the switch is released. However, there are also types with a toggle function of the RECORD switch. These are switched on when the RECORD switch is pressed, and off when the same switch is pressed again.

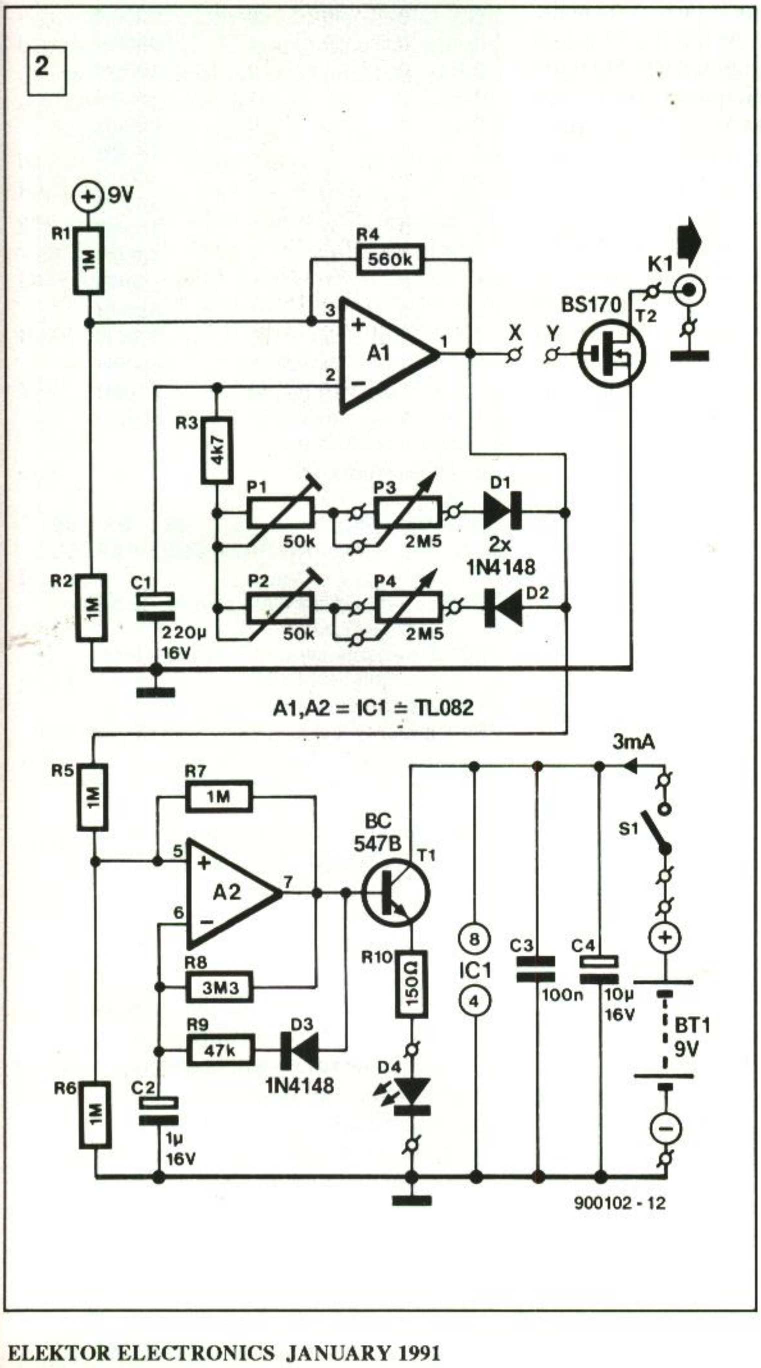
To allow the control to be used with the latter type of camcorder, the original circuit must be fitted with the extension shown in Fig. 1. The extension is inserted between terminals X and Y as indicated in Fig. 2. The track layout and component mounting plan in Fig. 3 show where to create terminals X and Y on the PCB.

With some camcorders, the proposed extension has one disadvantage: after switching on the camcorder, you are not sure whether it records or not when LED D4 flashes. If this problem occurs, make a habit of pressing the RECORD switch once, to make sure that the LED flashes when the camcorder is recording. ■

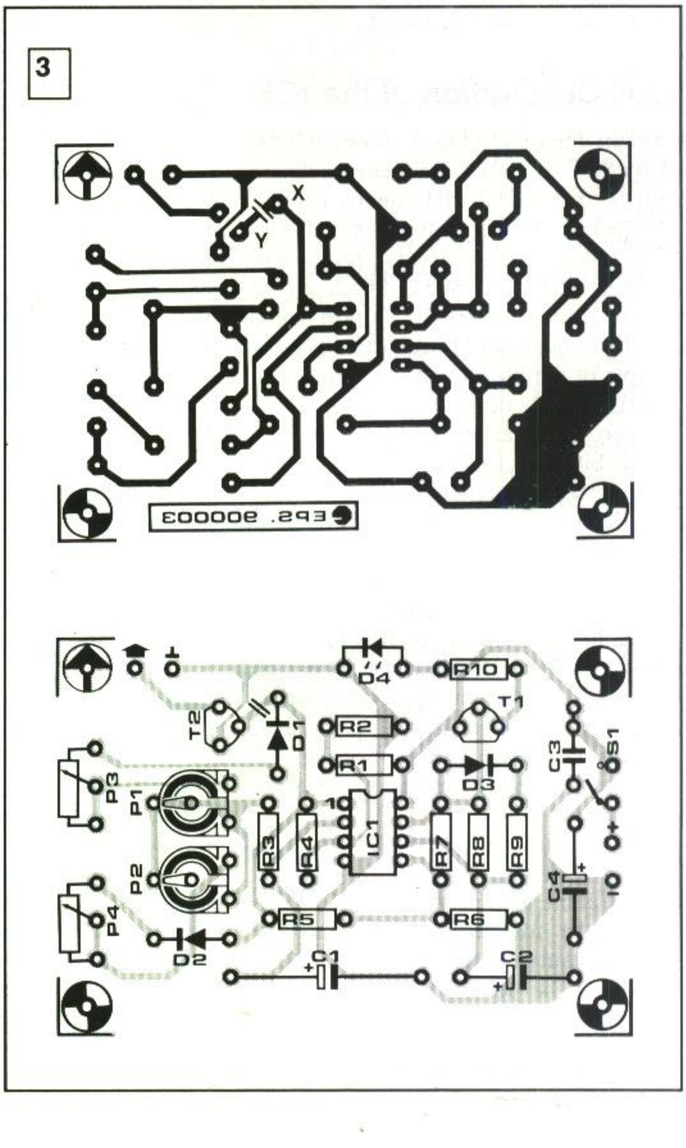
Reference:

"Interval control for camcorders", *Elektor Electronics* March 1990.

2



3



8031 IN-CIRCUIT EMULATOR

A previous article by the author on the 8031 microcontroller described some problems associated with developing single-chip microcontroller applications. The present article deals with the hardware that, when coupled with a PC and some software, enables the hobbyist to construct a low-cost in-circuit emulator (ICE) for the popular 8031 microcontroller.

Dr. David Kyte

THE hardware is based on the Dallas DS5000 family of processor. It has an 8031 architecture with battery-backed static RAM replacing the internal ROM. The actual processor is the Dallas DS2250, which is a SIP stick version of the DS5000. Functionally, the DS5000 and the DS2250 are identical. The DS2250, however, uses lower cost packaging techniques.

The hardware design is simple and constitutes the minimum required to build an operating DS2250 board. Consequently, it is a general-purpose software development board that will be used as an ICE.

Circuit description of the ICE

The circuit diagram in Fig. 1 shows that the clock for the DS2250 (IC1) is generated with the aid of an 11.0592 MHz quartz crystal. The frequency of the crystal may be familiar to those constructors who have previously built MCS-51-based (8031, 8052, 8751) computer boards with RS-232 port hardware. The frequency of the quartz crystal is a multiple of the bit rate on the serial port, i.e., 11.0592 MHz divided by the bit rate is an integer. This particular crystal enables the higher bit rates (19.2 KBit/s) to be generated. If a clock frequency of 12 MHz were chosen, the timers would be able to count intervals aligned more accurately to 1 s, but the user would be restricted to a maximum serial port speed of 4,800 baud.

Depending on its position, link J1 on the board allows the user to use the crystal on the ICE board, or that on the target board. If a link is installed between pins 1–3 and 2–4, the ICE crystal is used. If link positions 3–5 and 4–6 are selected, the crystal in the target system is used.

All of the pins on the DS2250 are connected to a 40-pin DIL socket, US1, which is positioned at the underside of the PCB. The board can be inserted directly into the 8031DIL socket on the target board. The processor can be powered either from the target board (power supplies on pin 40 (+5 V) and pin 20 (ground) of the DIL socket, or via header PL1. The power applied to header PL1 is 5 V on pin 2, and 0 V on pins 1 and/or 4. If the supply connector is accidentally reversed, no harm should come to the ICE.

Communications, initialization and downloading of software is provided via the RS-232 port on the PC. As the ICE51 uses a single 5-V supply, a combined RS-232 buffer/voltage converter is required. The device selected was the MAX232 (IC2) from Maxim. Four 10 µF capacitors, C4 to C7, are required for the voltage converter on board the MAX232. The RS-232 specification states that the high and low voltages must be more positive than +3 V and more negative than −3 V respectively. These voltages are provided by the MAX232 with typical values of +9 V and −9 V. A serial cable attached to PL2 connects the ICE51 and the PC serial port (PL2 is a 7×2 IDC connector). The 14-way connector was selected so that either a 9-way (AT-style) or a 25-way (XT-style) D-connector can be fitted at the PC end. Jumper J3 determines which D-type is in use. For an XT-style RS-232 connector, install links 7–6, 5–4 and 3–2. For an AT-style RS-232 connector, install links 7–8, 1–2 and 3–4.

Link J2 determines the source of the digital stream for the serial port to the receive pin on the processor. If J2 is connected in position 2–3, the data stream comes from the 8031 DIL header. In position 1–2, the data stream is connected to the output of the MAX232.

Downloading code to the ICE

The DS2250 has two modes of operation. The first mode executes code on reset from internal ROM. This ROM contains a small bootstrap loader program which is primarily designed to download code into the battery-backed RAM. Other operations include setting the memory configuration register, initialising the security keys and locking the RAM. To enter bootstrap mode, the $\overline{\text{PSEN}}$ line must be held low when the RST (reset) line is pulled high. The DS2250 then monitors the serial port, waiting for a 'carriage return' code, ASCII character 13. This operation is used to initialize the serial port baud rate. Thus, the sequence of actions to load a program into the DS2250 is:

Table 1. DS2250 register configurations

PA3	**PA2**	**PA1**	**PA0**	**Partition Address**
0	0	0	0	0000H
0	0	0	1	0800H
0	0	1	0	1000H
0	0	1	1	1800H
0	1	0	0	2000H
0	1	0	1	2800H
0	0	1	0	3000H
0	1	1	1	3800H
1	0	0	0	4000H
1	0	0	1	4800H
1	0	1	0	5000H
1	0	1	1	5800H
1	1	0	0	6000H
1	1	0	1	6800H
1	1	1	0	7000H
1	1	1	1	8000H

mcon register:

D7	**D6**	**D5**	**D4**	**D3**	**D2**	**D1**	**D0**
PA3	PA2	PA1	PA0	RA32	ECE2	PAA	SL

RA32 = embedded RAM size: 0 = 8 KByte; 1 = 32 KByte
ECE2 = enable second RAM chip (normally set to 0)
PAA = partition address access
SL = security lock

(1). Use a terminal emulator such as Procomm, or the ICE51 software.

(2). Toggle the reset switch, ensuring that it finishes in the 'on' position.

(3). Press the return key — the Dallas loader copyright message should then appear. You are now on line with the ICE.

(4). Unlock the external RAM using the 'U' command.

(5). Read the contents of the RAM configuration table by pressing 'R'. The configuration value after executing the 'U' command will be hexadecimal value F8. This register is used to define the size of the RAM available on the DS2250. Currently, the DS2250 supports three RAM sizes: 8 KBytes, 32 KBytes and 64 KBytes. The configuration register also determines the partitioning of the internal RAM into program and/or data memory. The internal RAM can be configured as data memory, program memory or split into part data and part program. The only constraint is that the RAM can not be both program and data memory within the same address range.

Table 1 describes the configuration permutations. When the ICE software is used, the memory configuration is normally set to $E8, which means that at least 2 KBytes of RAM from $7000 to $7FFF is configured as data memory, while RAM from $0000 to $6FFF is configured as program memory. **(6).** To write to the configuration register, use command 'W E8'.

The default setting for the 32-KByte ICE51 is hex E8. The partition address divides the battery-backed RAM into program memory (below the partition address) and data memory (above the partition address).

A program in Intel-hex format can now be loaded into the internal program memory. For instance, to load the ICE monitor program with the aid of the ICE51 host software, enter the load command, i.e., at the DOS prompt enter

DALLAS <CR>	Start the ICE51 software.
VDU <CR>	Enter terminal emulation mode. Set ICE51 switch on.
<CR>	Set the baud rate for the ICE51. Response should be the Dallas loader prompt.
L <CR>	Download Intel-hex fprmat file.
<PgDn>	Enter the file name to download.
DALLAS.OBJ <CR>	File name. Set the ICE51 switch S2 to off. Micro AMPS copyright message should appear.
<ESC>	Return to the host software; the monitor is now running.

Alternatively any terminal emulator, such as Procomm, that supports ASCII file transfers can be used.

At this point the monitor program has been downloaded, and the DS2250 has been configured correctly. To start executing the user code, switch S2 is set to 'off', and the user program will start executing from address $0000 as if the code had been masked in the internal ROM. The advantages of using internal ROM-ed code is that all four parallel ports are available for user applications. Two and a half of these ports would normally be required for the address, data and control bus when the processor executes in expanded mode.

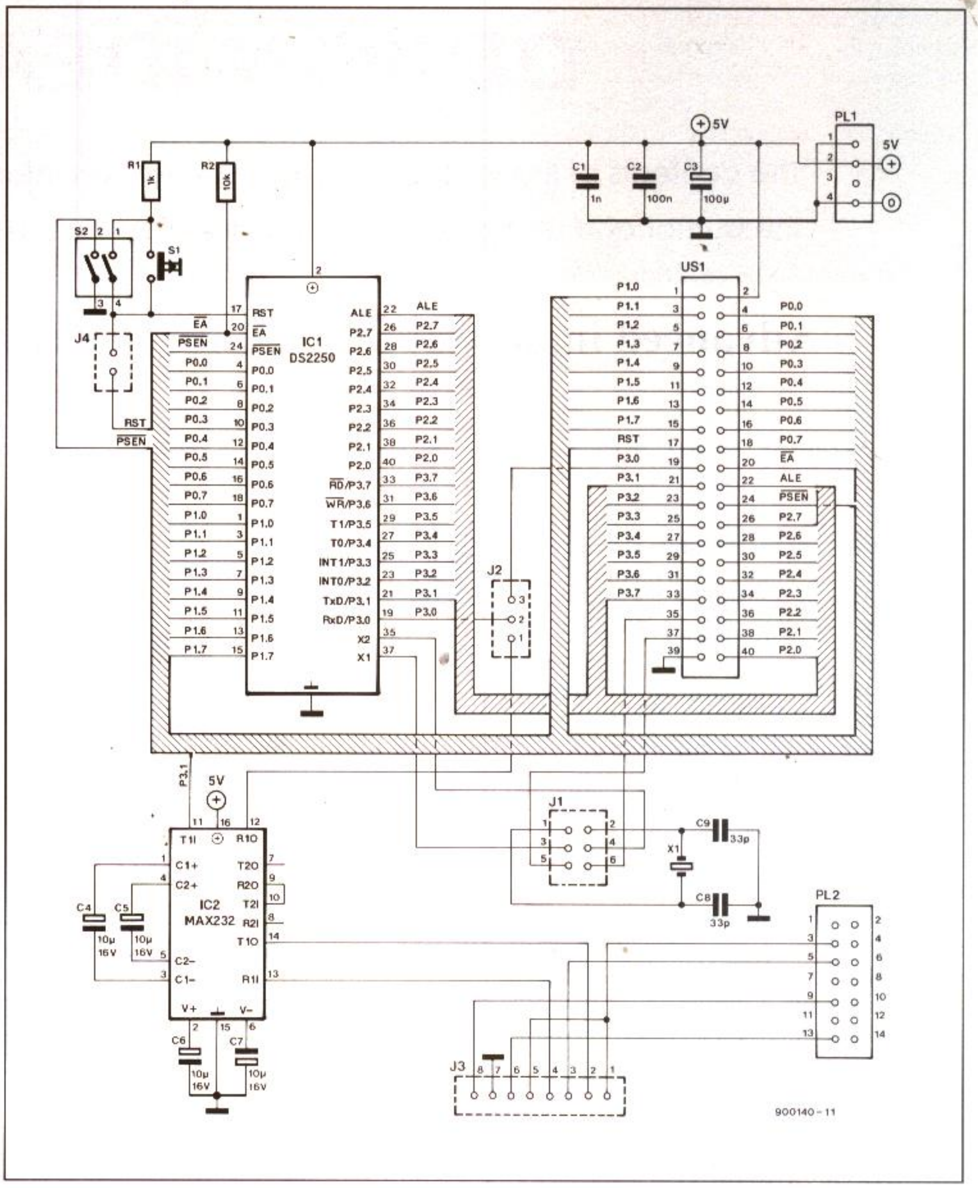


Fig. 1. Circuit diagram of the 8031 In-Circuit Emulator (ICE). Connector US1 is plugged into the socket for the 8031 in the target system. Software downloading is accomplished by means of a serial link to a PC running a terminal emulation program.

The ICE51 software package

The ICE51 software, which is available from Micro AMPS, provides debugging tools to convert the board from a DS2250 evaluation board into a full-blown single-chip mode 8031 in-circuit emulator. The 2 Kbytes large resident program is loaded into the DS2250 starting at address $0000 up to $07FF. User applications can then be loaded from $0800 onwards. The interrupt vectors are shifted by an amount of $800, i.e.,

Interrupt	**Old**	**New**
Ext. INT 0	$3	LJMP $803
Timer 0 overflow	$B	LJMP $80B
Ext. INT 1	$13	LJMP $813
etc.		

The software provides tools for reading and writing to internal and external memory, loading and saving Intel-hex code, reading and writing to the special function registers, disassembling code, starting programs, managing up to 10 breakpoints, and single-ep operation through programs. ■

Note:
The hardware and software described in this article are available from Micro AMPS Ltd. • 66 Smithbrook Kilns • Cranleigh • Surrey GU6 8JJ • ENGLAND. Telephone: (0483) 505395. Fax: (0483) 268397. The author is joint managing director of Micro AMPS Ltd.

Reference:

"The 8031/8751 microcontroller", *Elektor Electronics* July/August 1990.

DESIGN IDEAS

The contents of this column are based solely on information supplied by the author and do not imply practical experience by *Elektor Electronics*

Advanced input stage for a left/right revolution counter

by H. Matjaz

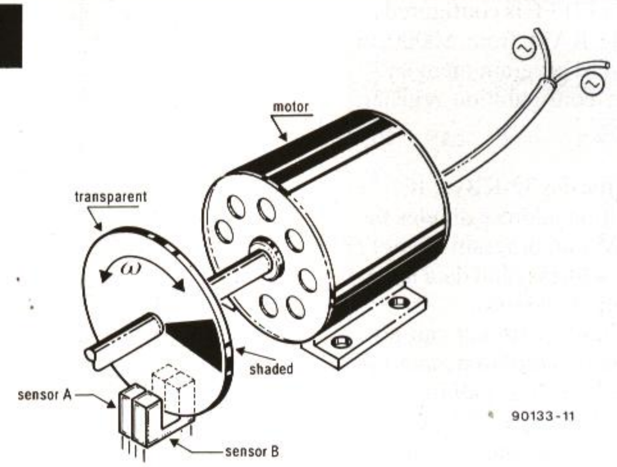
MOST left / right revolution counters are based on a shutter fixed on the axle of the motor that interrupts two infra-red sensors—see Fig. 1. Whether the counter counts up or down depends on which IR sensor is interrupted in the first place. Most constructors connect the IR sensors directly to the input of the counter. The minimum distance between the sensors is normally 5 mm.

A problem occurs when the motor stops at the exact moment the shutter is directly between the two sensors. The motor could then be turned the other way and the counter would count, although it should not. The problem arises, of course, because of the way the sensors are connected to the counter. Clearly, an input circuit is required that obviates this and similar problems. The one proposed here counts correctly, irrespective of how many times the motor changes direction ar at which angle it stops.

Method of counting

The method of counting is based on the traditional shutter and two IR sensors as shown in Fig. 1. The counter must count UP when the sensors are interrupted in the sequence shown in Fig. 2a. Logic high means that the sensor is interrupted; logic low that it is not interrupted.

Fig. 1



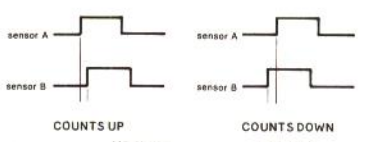


Fig. 2

Circuit description

The circuit of the input stage is shown in Fig. 3. When both sensors are not interrupted, the level at A and B is high (+5 V). When sensor A is interrupted, the level at A goes low. When sensor B is interrupted, the level at B goes low. Schmitt triggers Type

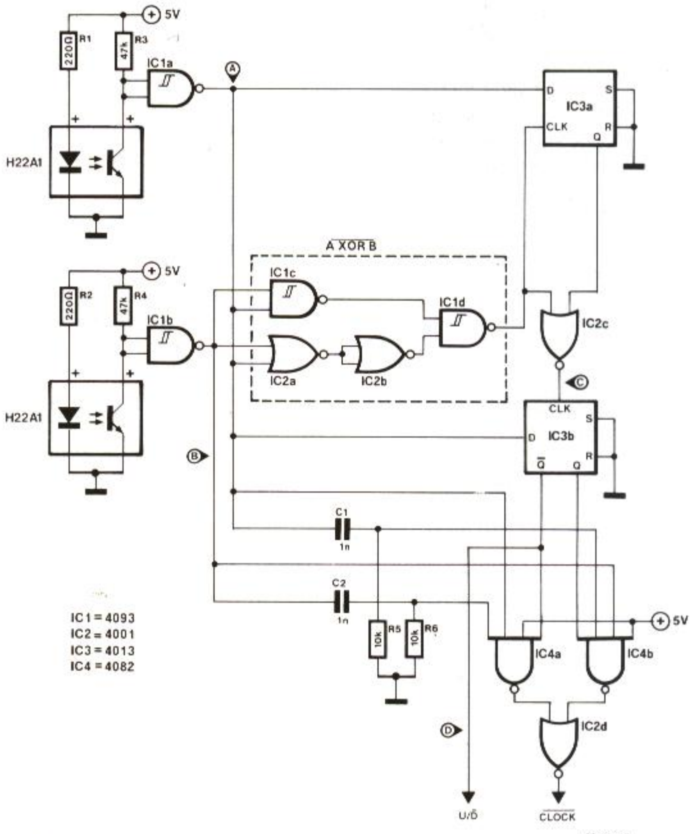


Fig. 3

Fig. 4

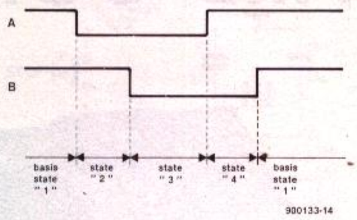


Fig. 5

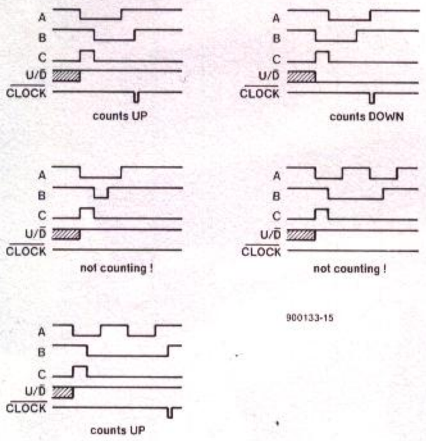
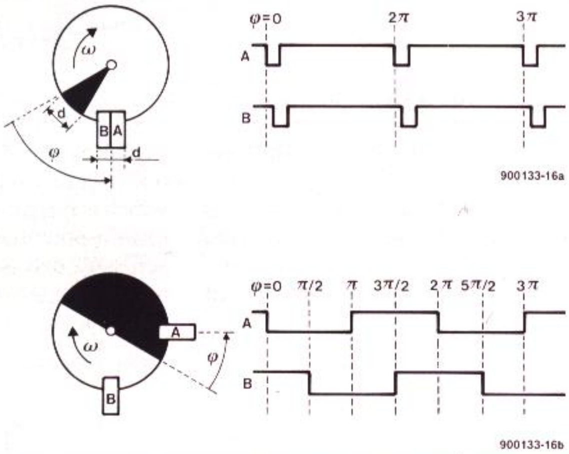


Fig. 6



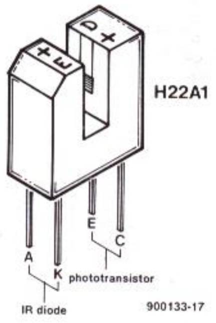
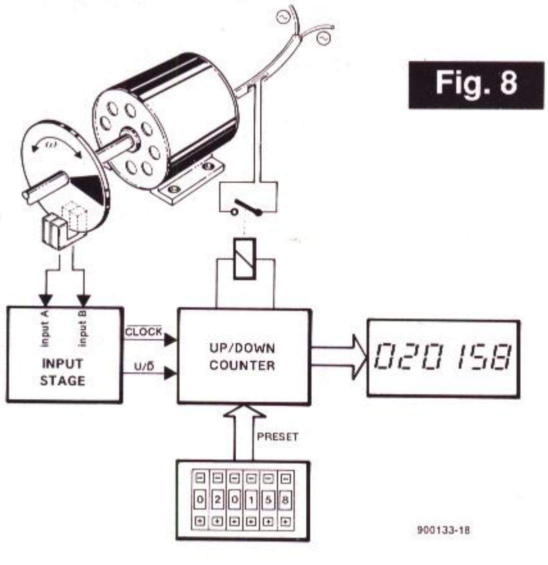


Fig. 7

Fig. 8



4093 are used here to ensure that the transition of the sensor output from high to low or vice versa is fast enough for the circuit to operate satisfactorily.

Production of an UP counting $\overline{\text{CLOCK}}$ pulse depends on three conditions being met:

- the level at D must be high;
- the level at A must be high;
- the level at B must change from low to high.

Similarly, production of a DOWN counting $\overline{\text{CLOCK}}$ pulse depends on:

- the level at D being low (that is, the Q output of IC_{3b} is high);
- the level at B being high;
- the level at A changing from low to high.

The four possible states are given diagrammatically in Fig. 4. Gates IC_{1c}, IC_{1d}, IC_{2a}, IC_{2b}, IC_{2c}, IC_{3a}, and IC_{3b} cause the level at D to be high only on the transition from state '1' to state '2'. The level at D is low on transition from state '1' to "4'. This means that the $\overline{\text{CLOCK}}$ for UP counting is produced only when sequential transition from state '1' to state '4' occurs (1–2–3–4–1). For DOWN counting, the transition should be 1–4–3–2–1. Some examples of when counting takes place or not are given in Fig. 5. Time constants R_5–C_1 and R_6–C_2 determine the width of the CLOCK pulse: for instance, if $R_5 = R_6 = 10\ \text{k}\Omega$ and $C_1 = C_2 = 1\ \text{nF}$, the width is $10\ \mu\text{s}$.

When $U/\overline{D}$ is high, the counter must count UP, otherwise DOWN.

The shutter is made of transparent material with a sector blacked as shown in Fig. 6. The sector may have various areas, while the sensors may be placed in a number of locations. Two will be discussed.

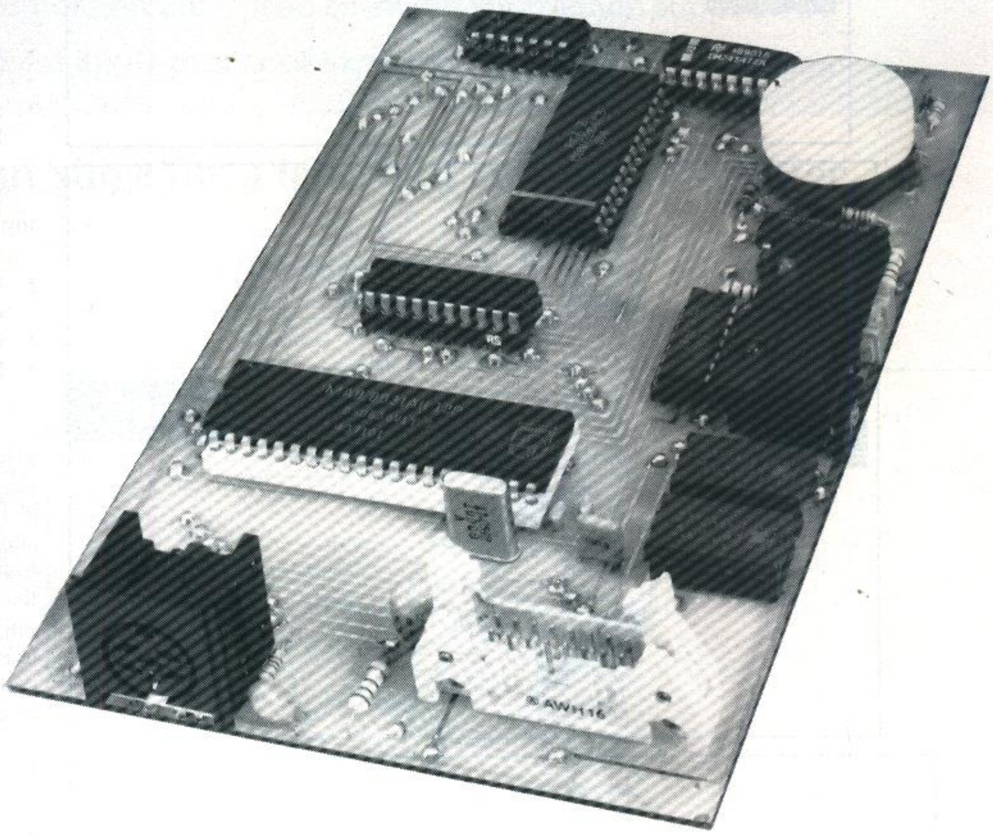
The size of the shutter sector and the position of the two sensors in Fig. 6a is suitable for devices where the counting UP must be at the same angle as the counting DOWN ($\phi = 0$). This arrangement would find application in, for instance, a transformer winding system.

The size of the sector and the sensor positions shown in Fig. 6b are suitable for applications where the motor has a high angular velocity.

The construction of a typical IR sensor, the Type H22A1, is shown in Fig. 7, while a typical application of the input stage is given in Fig. 8. ■

8031 SINGLE BOARD COMPUTER

Intel's 8031 embedded controller chip contains all the essential ingredients of a microcomputer — I/O ports, a UART, interrupt lines, 16-bit timers and counters, an 8-bit CPU, and 128 bytes of RAMstack. All that is additionally required to form a fully functional computer is an operating system, a means of getting machine code to the CPU, a few operating switches and a visual display. Such a computer is described here.



R. Grodzik

THIS project uses the Intel 8031 microcontroller to form a self-contained, machine code programmable computer board, measuring only 100 × 160 mm. The operating system, all 240 bytes of it (probably the world's smallest...), is contained in a preprogrammed PROM. The project is aimed in particular at experienced software and hardware designers who want to use the 8031 in stand-alone application circuits, such as automated control.

Circuit description

The block diagram of the single board computer, Fig. 1, shows the main elements of the computer: the 8031 MPU (master processing unit), 8 KBytes of RAM (random-access memory), a PROM (programmable read-only memory) and a data indication based on two LED displays. A handful of logic ICs complete the computer.

Port 1 (lines P1.0–P1.7) provides eight I/O lines, (latched input or output), whose logic status is continuously monitored by an intelligent hexadecimal display. Port 3 (lines P3.1–P3.4) provides four additional I/O lines, of which three have a dual function of external interrupt and counter input.

The computer is controlled by the operating system firmware resident in IC3, a PROM. The instructions in this PROM enable the 8031 to read the data from the serial input port, and send it to the RAM. In addition, the PROM provides the interrupt vectors for the system.

A major and useful feature of the board is the provision of battery back-up for the 6264 static RAM. As shown in the circuit diagram, Fig. 2, the RAM supply switching circuit consists of IC6, R6, R7, R8 and D3. With a 5-volt supply present on the board, the 3.6-V NiCd battery is permanently charged via D3, which is forward biased. Resistors R7 and R8 form a potential divider. The voltage drop across R8 drives transistor T2, whose collector voltage is brought low. Via the output of IC6B, a logic low is applied to pin 1 of IC6A.

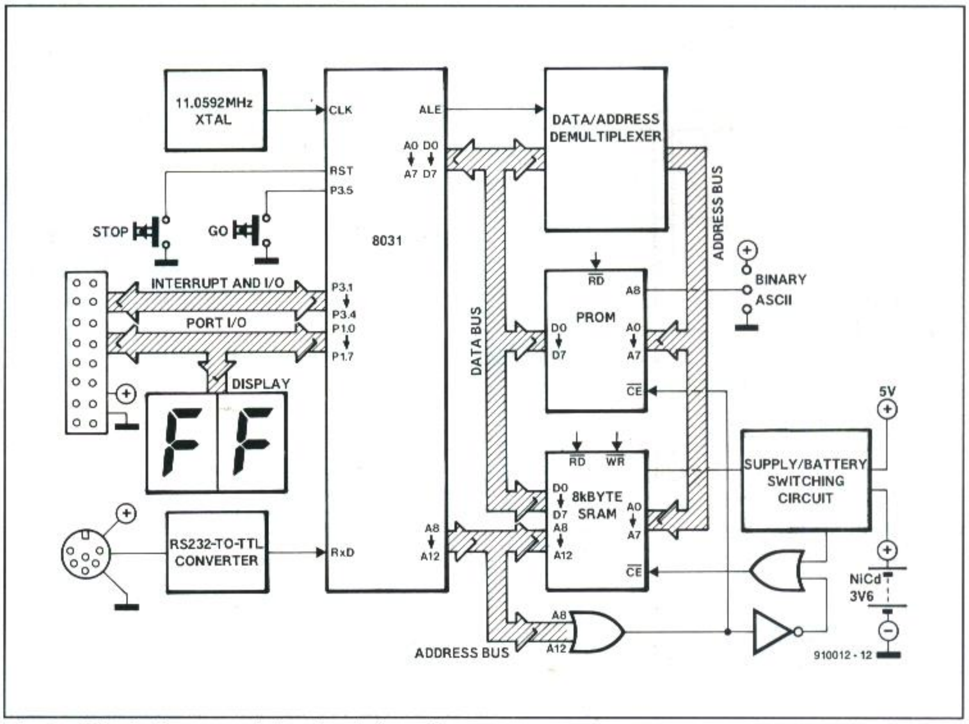


Fig. 1. Block diagram of the single-board computer.

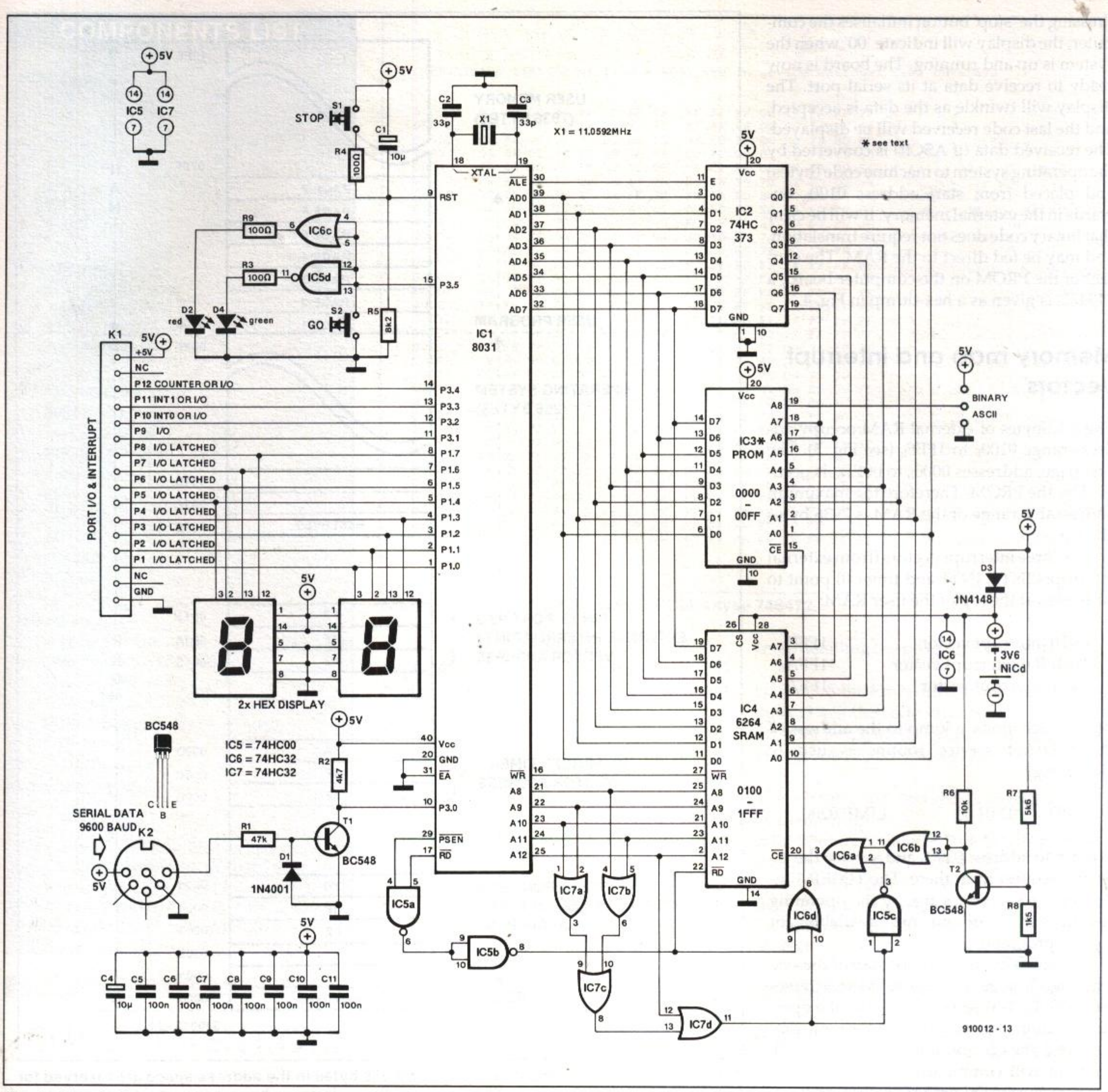


Fig. 2. Circuit diagram of the 8031-based computer. Data downloaded from the PC is shown on two 'intelligent' hex diplay devices connected to port 1 of the 8031.

Pin 2 of IC6A remains high when the PROM is addressed. Above address $00FF_h$, one or more of the inputs of IC7, configured as a five-input OR-gate, go high, which in turn is inverted by IC5C. Next, a 'low' is applied to pin 2 of IC6A; pin 3 of IC6A goes low and thus enables RAM IC4, while at the same IC3 is disabled.

When power is removed from the board, diode D3 becomes reversed biased, preventing any current from the battery reaching the base of T2. Consequently, this transistor remains off, with its collector at battery voltage. The output of buffer IC6B (pin 11), is therefore high. This propagates through IC6A to pin 20 of IC4, thus disabling the RAM. Simultaneously, keep-alive voltage is supplied to pins 26 and 28 of IC4 to retain the memory contents.

Programming the computer

These days, most people have access to a personal computer (PC), and with the addition of one of many communications software utilities available, or, indeed, an 8031 assembler, it is relatively simple to produce machine code and download it via the PC's serial port to the input of the UART contained in the 8031. The code then resides in the on-board 8 KByte-large battery-backed RAM, which ensures that the downloaded program will remain intact in the event of loss of power. This method of programming a computer is particularly attractive because it dispenses with the time-consuming practice of programming and erasing EPROMs whilst debugging the software. Here, it is a simple matter to re-edit the code on the host PC, and then download the final version to the computer board at 9600 baud. Even for the full eight kilobytes, the transfer only takes a few seconds.

The author has deliberately avoided the use of Intel-hex or Motorola-s protocols. Object code can be downloaded in binary or ASCII format. The binary code is accommodated as byte-wide RS232 transmissions: 8 data bits, 1 stop bit, no parity, 9600 bits per second. Alternatively, each byte can be downloaded as two ASCII words: i.e., $2F_h = 32_{ASCII}$ followed by 46_{ASCII}. Line feeds, return codes and spaces must not be used in this format because the board's operating system will identify them as invalid codes, and the display will start blinking rapidly.

The board control switches could not be simpler: one 'go' and one 'stop' push-button.

Pressing the 'stop' button initialises the computer; the display will indicate '00' when the system is up and running. The board is now ready to receive data at its serial port. The display will twinkle as the data is accepted, and the last code received will be displayed. The received data (if ASCII) is converted by the operating system to machine code (bytes) and placed from start address 0100_h onwards in the external memory. It will be clear that binary code does not require translation, and may be fed direct to the RAM. The content of the PROM on the computer board, a 74S472, is given as a hex-dump in Fig. 4.

Memory map and interrupt vectors

The 8 KBbytes of external RAM occupy address range 0100_h to $1FFF_h$ (see Fig. 3). The first page, addresses 0000_h to $00FF_h$, is occupied by the PROM. Therefore the maximum addressable range of the RAM is 7936 bytes ($1F00_h$).

The three interrupt vectors (from external interrupts INT0, INT1 and timer 0) point to addresses at the top of the user RAM:

INT0 interrupt vector	$1FFD_h$
TIMER 0 interrupt vector	$1FFA_h$
INT1 interrupt vector	$1FF7_h$

At these locations, a jump to the address of the interrupt service routine is usually placed, e.g.:

```
1FFD:   02 02 00          LJMP 0200
```

to jump to address 0200_h and service the interrupt request from there. The TIMER 1 interrupt is reserved for use by the operating system and therefore not available for general programming.

As shown by the memory map of the system, Fig. 3, page 0 of the RAM (addresses 0000_h-$00FF_h$) is reserved for use by the operating system. To execute the program, simply press the 'go' button. Execution of the program will commence at address 0100_h. Connect on an ADC or DAC to the port lines, and there you have it: computer control of the outside world. Data acquisition, monitoring, counting, timing — the list is endless; it all depends on you.

Construction

Construction of the 8031 computer is fairly simple, as relatively few parts are involved. A double sided PCB is used with special purpose 0.8-mm pins for pinning through. These pins are much better than odd bits of thin wire which tend to break or fall through. The CMOS displays are static sensitive and fairly expensive, so use the standard static precautions when handling these devices. It is recommended to fit the displays and all ICs in sockets, which make for easier fault-finding. The NiCd battery is secured to the PCB with a few drops of two-component glue.

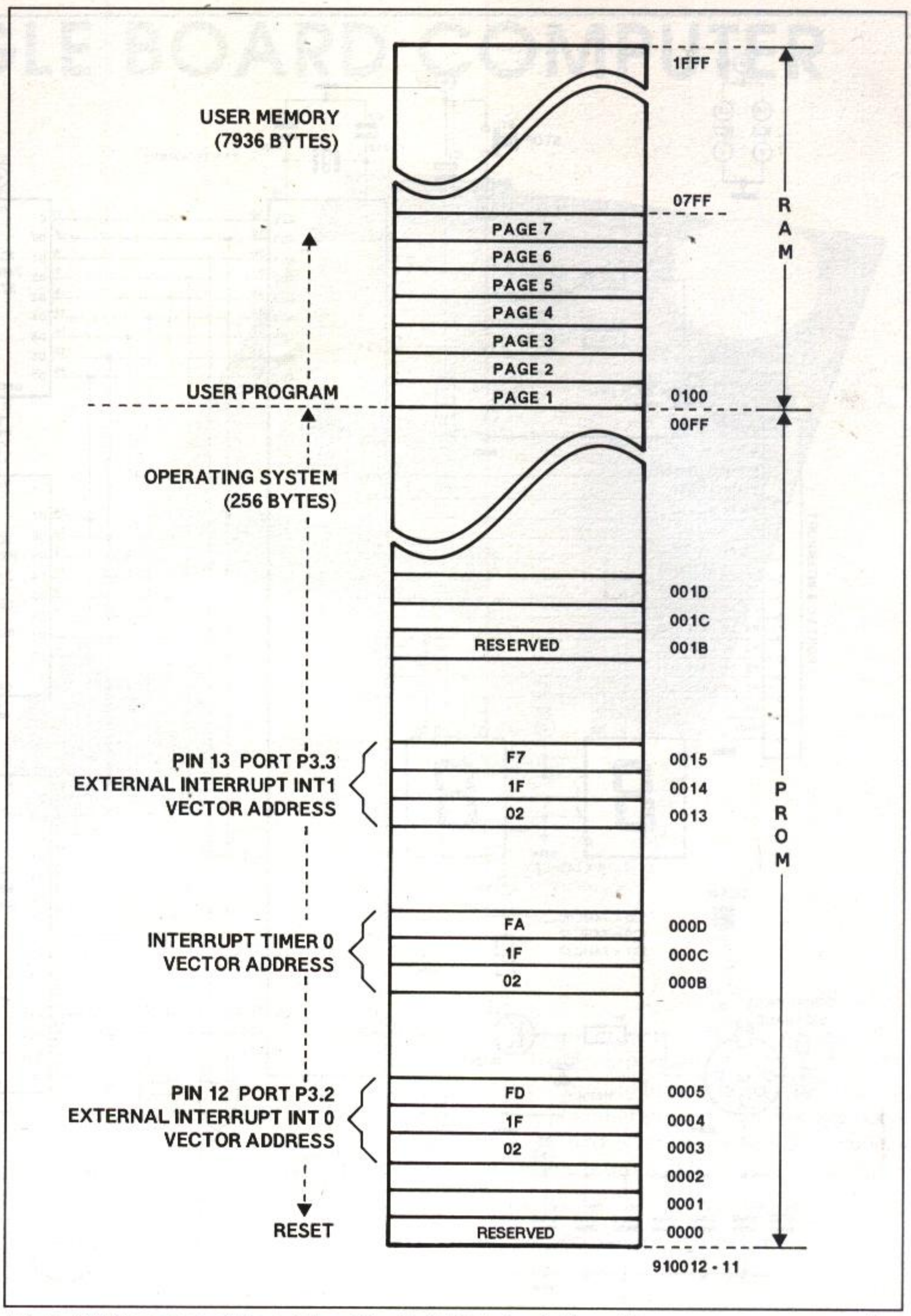


Fig. 3. Address map of the system. The first 256 bytes in the address space are reserved for the firmware PROM that contains the operating system and a number of interrupt vectors. The user memory has a size of slightly less than eight kilobytes.

For binary download operation, strap pin 19 of IC3 to +5 V (pin 20); for ASCII downloading, strap pin 19 of IC3 to 0 V (pin 10).

Once the discrete components, the crystal and IC sockets have all been soldered into place, check for solder bridges and 'dry joints'. Use an ohmmeter to check for continuity between all parts of the circuit. Next, apply power (5 V), and with the aid of a logic probe, ensure that the supply voltages and 0 V potentials are present at the appropriate pins of each IC socket. For example, check IC2 pin 20 (+5 V), and pin 1 and pin 10 (both 0 V).

Disconnect the power and insert IC1, the 8031, into its socket. This is where the author's fault-finding tool, an inexpensive transistor radio, comes in. Power up the board, tune into the medium-wave band and you should hear a hum of microprocessor activity. Placing your finger on the PROM socket to simulate software will modulate the output from the radio. If this does not happen, proceed no further, as there is probably an open or short on your circuit board. Investigate and rectify. Next, switch off and populate the board with the remaining ICs. Power up and check with a logic probe that pins 32 up to and including 39 of IC1 are strobing. If everything is all right, the display reads '00'. If not, recheck everything.

Connecting up

Data connections to the serial port are via a 6-way DIN socket and mating plug. Note that the board is also powered via this DIN

COMPONENTS LIST

Resistors:

1	47K	R1
1	4K7	R2
3	100Ω	R3;R4;R9
1	8kΩ2	R5
1	10kΩ	R6
1	5kΩ6	R7
1	1kΩ5	R8

Capacitors:

2	10µF 16V tantalum bead	C1;C4
2	33pF	C2;C3
7	100nF ceramic	C5 – C11

Semiconductors:

1	8031 or 80C31	IC1
1	74HC373	IC2
1	PROM 74S472 (see note)	IC3
1	6264	IC4
1	74HC00	IC5
2	74HC32	IC6;IC7
2	1N4148	D1;D3
2	BC548	T1;T2
2	LED (1 green, 1 red)	D2;D4

Miscellaneous:

2	keyboard rocker switch with indicator (RS 319-843)	S1;S2
1	11.0592 MHz quartz crystal	X1
1	16-way IDC box header	K1
1	PCB-mounting 6-way DIN socket	K2
1	3V6 NiCd battery	
0.8-mm pcb pins		

Note:
The pre-programmed PROM and the PCB artwork for this project are available from:
R. Grodzik • 53 Chelmsford Road • BRADFORD BD3 8QN • ENGLAND. The price of the PROM is £15.00 including P&P.

```
SPRINTplus listing of file A:8031.prm in device National Semi DM74S472

00000:  02 00 C0 02 1F FD FF FF  FF FF FF 02 1F FA FF FF   '................'
00010:  FF FF FF 02 1F F7 FF FF  FF FF FF 02 00 3A FF FF   '.............:..'
00020:  C2 F0 90 01 00 D2 B5 75  98 50 75 89 20 75 8D FD   '.......u.Pu. u..'
00030:  D2 8E D2 AF D2 AB 80 FE  80 FC 20 B5 09 75 09 00   '.......... ..u..'
00040:  75 08 D0 C2 8E 32 30 98  F1 C2 8E C2 98 E5 99 54   'u....20........T'
00050:  7F F5 90 12 00 68 10 F0  08 C4 F5 70 D2 F0 D2 8E   '.....h.....p....'
00060:  32 25 70 F0 A3 D2 8E 32  F8 54 F0 B4 30 02 80 18   '2%p....2.T..0...'
00070:  B4 40 06 C3 E8 94 47 40  02 80 13 E8 B4 40 02 80   '.@....G@.....@..'
00080:  F8 C3 94 37 22 00 00 00  E8 C3 94 3A 40 02 80 0B   '...7"......:@...'
00090:  E8 94 2F 40 06 C3 E8 94  30 80 E9 74 00 F5 90 12   '../@....0..t....'
000A0:  00 AB 74 88 F5 90 12 00  AB 80 F0 7E 80 1E 7F FF   '..t........~....'
000B0:  1F BF 00 FC BE 00 F6 22  FF FF FF FF FF FF FF FF   '......."........'
000C0:  74 00 F5 90 02 00 20 FF  FF FF FF FF FF FF FF FF   't..... .........'
000D0:  12 00 AB C2 B5 02 01 00  FF FF FF FF FF FF FF FF   '................'
000E0:  FF FF FF FF FF FF FF FF  FF FF FF FF FF FF FF FF   '................'
000F0:  FF FF FF FF FF FF FF FF  FF FF FF FF FF FF FF FF   '................'
00100:  02 00 6B 02 1F FD FF FF  FF FF FF 02 1F FA FF FF   '..k.............'
00110:  FF FF FF 02 1F F7 FF FF  FF FF FF 02 00 3A FF FF   '.............:..'
00120:  C2 F0 90 01 00 D2 B5 75  98 50 75 89 20 75 8D FD   '.......u.Pu. u..'
00130:  D2 8E D2 AF D2 AB 80 FE  80 FC 20 B5 09 75 09 00   '.......... ..u..'
00140:  75 08 63 C2 8E 32 30 98  F1 C2 8E C2 98 E5 99 F5   'u.c..20.........'
00150:  90 F0 A3 D2 8E 32 7E 80  1E 7F FF 1F BF 00 FC BE   '.....2~.........'
00160:  00 F6 22 12 00 56 C2 B5  02 01 00 74 00 F5 90 02   '.."..V.....t....'
00170:  00 20 FF FF FF FF FF FF  FF FF FF FF FF FF FF FF   '. ..............'
00180:  FF FF FF FF FF FF FF FF  FF FF FF FF FF FF FF FF   '................'
00190:  FF FF FF FF FF FF FF FF  FF FF FF FF FF FF FF FF   '................'
001A0:  FF FF FF FF FF FF FF FF  FF FF FF FF FF FF FF FF   '................'
001B0:  FF FF FF FF FF FF FF FF  FF FF FF FF FF FF FF FF   '................'
001C0:  FF FF FF FF FF FF FF FF  FF FF FF FF FF FF FF FF   '................'
001D0:  FF FF FF FF FF FF FF FF  FF FF FF FF FF FF FF FF   '................'
001E0:  FF FF FF FF FF FF FF FF  FF FF FF FF FF FF FF FF   '................'
001F0:  FF FF FF FF FF FF FF FF  FF FF FF FF FF FF FF FF   '................'
```

910012 - 14

Fig. 4. Hexdump of the firmware PROM, a type 74S472.

socket. At the host computer end, connect a single screened cable to the ground and TXD pins of the serial port. Also tie the RTS and CTS pins together.

All the port I/O, interrupt, and power lines of the 8031 computer are brought out to a 16-way IDC pin header (K1), for connection via ribbon cable to peripheral devices. As shown in the photograph of the prototype, the pin header is a box type connector with eject headers.

For further reading

Various publications are available from Intel, detailing the programming of the 8031:

(1) Intel 8-bit Embedded Controller Handbook (1989) (RS code 910-749);

(2) Intel Embedded Controller Applications Handbook (RS code 910-777);

These books, and data sheets on the 8031, are available from

Intel Literature Sales • P.O. Box 7641 • Mt. Prospect • IL 60056-7641 • USA.
or, in the UK, from
Intel Corporation (UK) Ltd. • Pipers Way • Swindon • Wilts SN3 1RJ. Telephone: (0793) 696000. ■

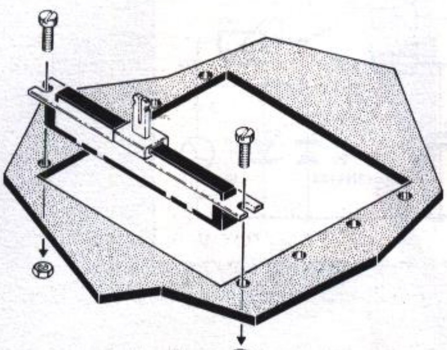
SLIDE POTENTIOMETERS IN THE VIDEO MIXER — AN UPDATE

We understand that the mounting of the slide potentiometers in the video mixer published last year has caused a small difficulty with some constructors.

There appear to be two types of slide potentiometer around, which, although they have the same track length, are mounted differently. In some cases, the type with two mounting lugs (Fig. 1) requires a few washers, or short PCB spacers, to be positioned at the right height above the PCB. The second type (Fig. 2) has two holes through the potentiometer body. To enable this type to be secured to the PCB, mount two small support plates and two spacers at the track side of the PCB, as shown in Fig. 2. The length of the spacers is determined by the required height of the slide potentiometer above the PCB surface. ■

"Video Mixer", *Elektor Electronics* January, February and March 1990.

1



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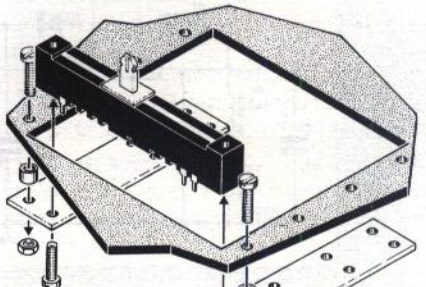


SWR METER FOR 1.5 – 70 MHz

Waste not, want not! Radio amateurs are a breed of electronics enthusiasts who like to see every milliwatt of their precious RF power arriving where it belongs: at the antenna terminals. However, budgets being what they are (RF plugs and low-loss coax cable are pretty expensive…), impedance mismatches occur readily between the transmitter and the aerial, causing power reflection. The low-cost instrument described here will tell hams operating in the 160 m through 4 m bands the ratio of the forward to the reflected RF power.

J. Bareford

A standing-wave ratio (SWR) meter is an indispensable RF power monitoring instrument found in almost any radio amateur's shack. It is often connected permanently between the transmitter/receiver rig and the coax cable to the antenna. During transmissions, it provides a relative indication of the transmitted power, as well as an indication of the ration of the forward RF power (i.e., the power fed to the antenna) to the reflected RF power (i.e., the power reflected into the transmitter owing to a mismatch at some point in the transmission line).

Since most radio amateurs have several transmitters and antenna systems, it is good operating practice to check the antenna match at a low power level, before starting a transmission. In this way, hams protect their precious RF amplifiers from going up into smoke owing to a gross mismatch, a faulty coax relay, or a totally absent antenna connection. During the transmission, hams have one eye on the tuning scale, and the other on the needle of the moving-coil meter in the SWR instrument. Just for reassurance? No, a must to ensure the best possible signal at the receiving station.

Standing-wave ratio

The ratio of the forward RF power to the reflected RF power is called the standing-wave ratio, or SWR. Since, in the actual SWR meter, we are working with one, definitely established, transmission line impedance, the term 'power' may be replaced by 'voltage'. Thus, in a transmission line system where a mismatch exists, we have a forward voltage, U_F, and a reflected voltage, U_R. This allows the SWR to be determined from

$$\mathrm{SWR} = \frac{U_F + U_R}{U_F - U_R}$$

From this, it is seen that an SWR of 1 corresponds to optimum matching of the transmitter to the coax and antenna system. That is not to say that there are no losses: it only means that the transmitter output is matched to the load impedance formed by the transmission line, which includes the aerial, i.e., the load at the far end of the cable. In fact, the effect of an improperly matched antenna becomes smaller as the attenuation on the cable rises. This is because both the forward and the reflected power are subject to the same attenuation. A good discussion of the relative importance of the SWR is found in Ref. 1. Incidentally, long runs of lossy coax cable often form the perfect dummy load: any length of (matched) coax cable that introduces an attenuation of more than, say, 20 dB at the test frequency, will reflect so little power that it looks like a perfect resistance to the transmitter (keep an eye on the dissipation, though!).

Typical, tolerable, SWR values resulting from small mismatches are in the range from 1.5 to 2.0. Most radio amateurs would agree that an SWR greater than 2 is a definite cause for investigating the mismatch, as then more than 11% of the transmitted power is 'wasted' by reflection. The cause of the mismatch may be found in an incorrectly tuned RF amplifier, a piece of coax with the wrong impedance (the notorious 50/75-Ω prob-

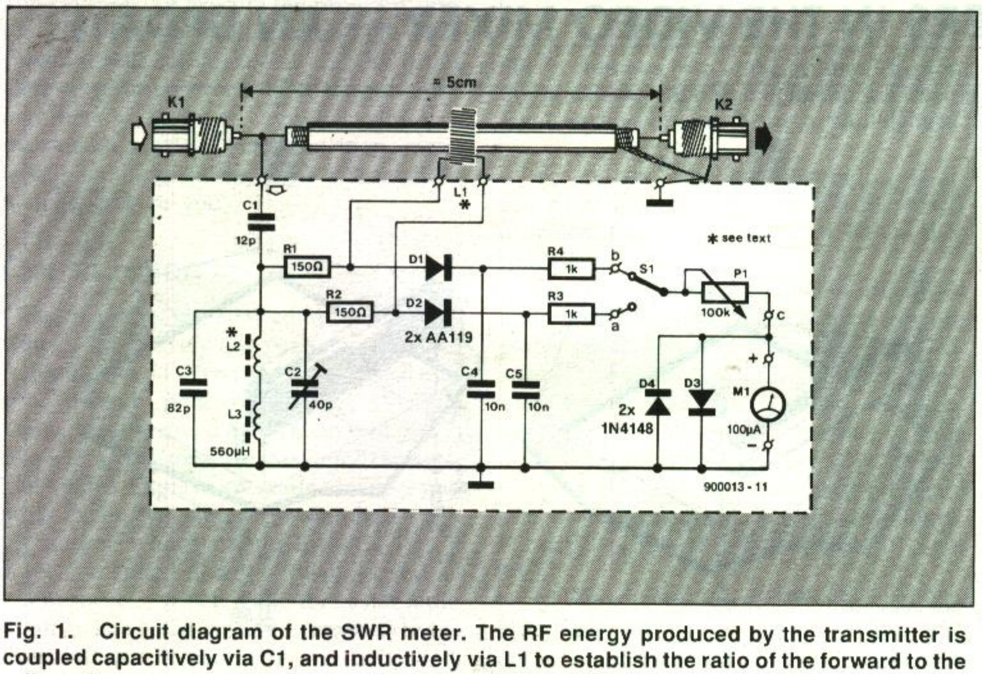


Fig. 1. Circuit diagram of the SWR meter. The RF energy produced by the transmitter is coupled capacitively via C1, and inductively via L1 to establish the ratio of the forward to the reflected power.

lem), birds on your aerial, a frozen-over aerial, the hectic of a contest, water in your coax cable, or a couple of faulty RF connectors somewhere on your attic or on the roof. Whatever the cause, make sure you eleminate it before starting to transmit, since many RF amplifiers, particularly those used for SSB (single-sideband) and other linear modes, do not like output mismatches, and produce an excessively wide output spectrum, causing splatter in the band and annoyance to your fellow hams in the neighbourhood.

The circuit

The circuit diagram of the present SWR meter is conventional, see Fig. 1. The forward and reflected powers induce RF voltages in a toroid inductor, L1, which is positioned around a short length of coax cable. Note that the cable is grounded at one side of the instrument only (a coax cable grounded at both ends does not radiate).

The RF voltage supplied by the transmitter is capacitively coupled via C1 to serve as a kind of reference against which the forward and reverse powers are measured. The coupling capacitor is connected to a tuned circuit, L2-L3-C3-C2, that serves to balance the measurement circuit at higher frequencies (in the 6 m band and possibly the 4 m band also).

The forward and reflected voltages are rectified by two diodes, D1 and D2, to establish the relative powers and thus the SWR. The AA119s used are low-capacitance point-contact germanium diodes with a low threshold voltage of about 0.2 V. A toggle switch, S1, allows the user to select a (relative) forward power indication, or reflected power relative to forward power.

Construction

The layout of the double-sided printed circuit board is shown in Fig. 2. Be sure to avoid overheating the trimmer, C2, while soldering its terminals. As shown in the photograph in Fig. 3, the two BNC sockets are connected by a short length of thin 50-Ω coaxial cable, of which the screening braid is connected to the socket and the board **at the antenna side only**. In the prototype, the coax cable was a 40-mm long piece of RG174U, which has an outside diameter of about 3 mm.

The winding data of the two inductors in the instrument, L1 and L2, are as follows:

Pick-up inductor L1:

Wind 30 turns of 0.2-mm dia (SWG36; AWG34) enamelled copper wire on a FT 37-43 ferrite ring core from Amidon Associates Inc. An alternative core is the FB 43-2401. Distribute the wire evenly on the core. Carefully remove the enamel coating at the ends. Put the coax cable through the hole in the core, and solder the wire ends of the inductor to the holes marked 'L1' on the PCB. Connect the coax cable to the BNC sockets as indicated above.

Choke L2:

This is made from 6 turns of 0.2 mm dia. (SWG36; AWG34) enamelled copper wire through a 3-mm long ferrite bead. After winding the inductor, carefully remove the enamel coating at the ends, and solder the device in place.

The completed printed circuit board fits in an Eddystone or Hammond diecast enclosure of about 11×6×3 cm. The size of the rectangular clearance in the lid of the enclosure depends on the meter you use. The lid is drilled to accept the threaded shafts of the sensitivity control potentiometer and the forward/reflected power switch. The connec-

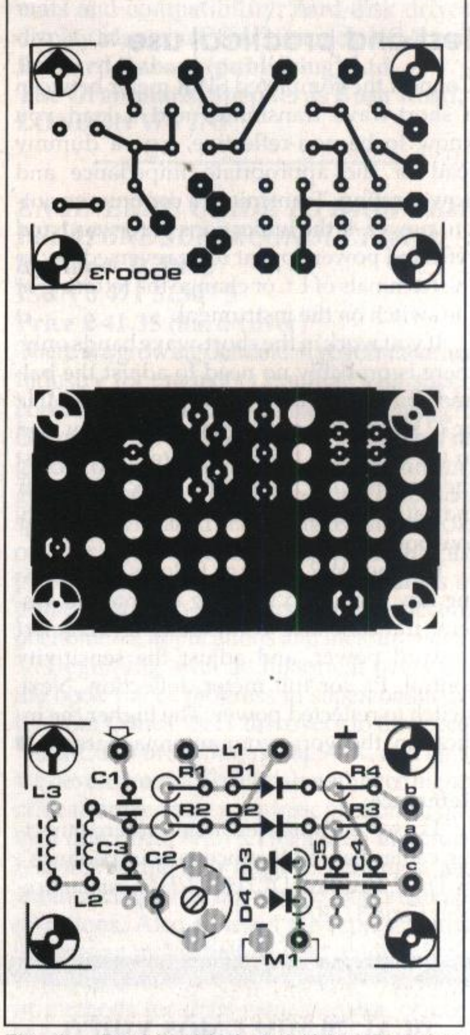


Fig. 2. Double-sided, not through-plated, printed circuit board for the SWR meter.

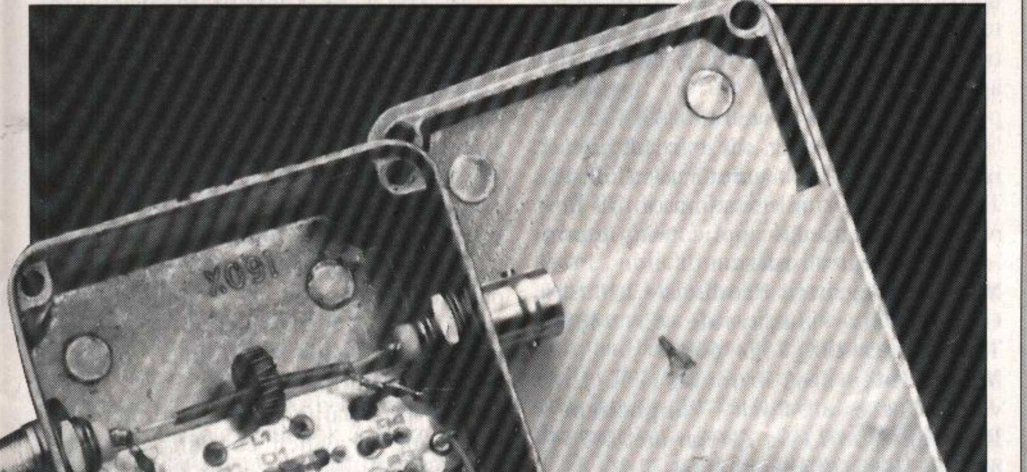


Fig. 3. A look inside the prototype. Note that the toroid core, L1, is fitted around the 3-mm dia. coax cable connected between the input and output BNC sockets.

COMPONENTS LIST

Resistors:

2	150Ω	R1;R2
2	1kΩ	R3;R4
1	100kΩ linear potentiometer	P1

Capacitors:

1	12pF	C1
1	40pF trimmer	C1
1	82pF	C3
2	10nF	C4;C5

Semiconductors:

2	AA119	D1;D2
2	1N4148	D3;D4

Inductors (see text):

1	FT37-43 (Amidon)	L1
1	3-mm ferrite bead	L2
1	560µH	L3

enamelled copper wire 0.2 mm dia.

Miscellaneous:

1	miniature SPDT switch	S1
1	100µA moving-coil meter	M1
2	BNC socket	K1;K2
1	diecast enclosure, e.g., Hammond type 1590B	

tions between the board and the external components are shown in Fig. 4.

Test and practical use

Connect the completed SWR meter between a short-wave transmitter and a load you know to be non-reflective, e.g., a dummy load of the appropriate impedance and power rating. Transmit at a continuous output power. If the indications of forward and reflected power appear to be reversed, swap the terminals of L1, or change the lettering of the switch on the instrument.

If you work in the short-wave bands only, there is probably no need to adjust the balance trimmer, C2. If you do work at 50 MHz or 72 MHz (UK only), connect a dummy load to the output of the SWR meter, and adjust the trimmer for minimum reflected power indication while transmitting at a relatively low power.

Use the SWR meter as follows after making any change to your rig or antenna system: transmit at a continuous level, select forward power, and adjust the sensitivity control, P1, for full meter deflection. Next, switch to reflected power. The higher the indication, the worse your antenna match. ■

Reference:

1. "Losses encountered when interconnecting cables having the incorrect impedance". by Dr. P. Brumm, DL7HG. *VHF Communications* issue 3/1974.

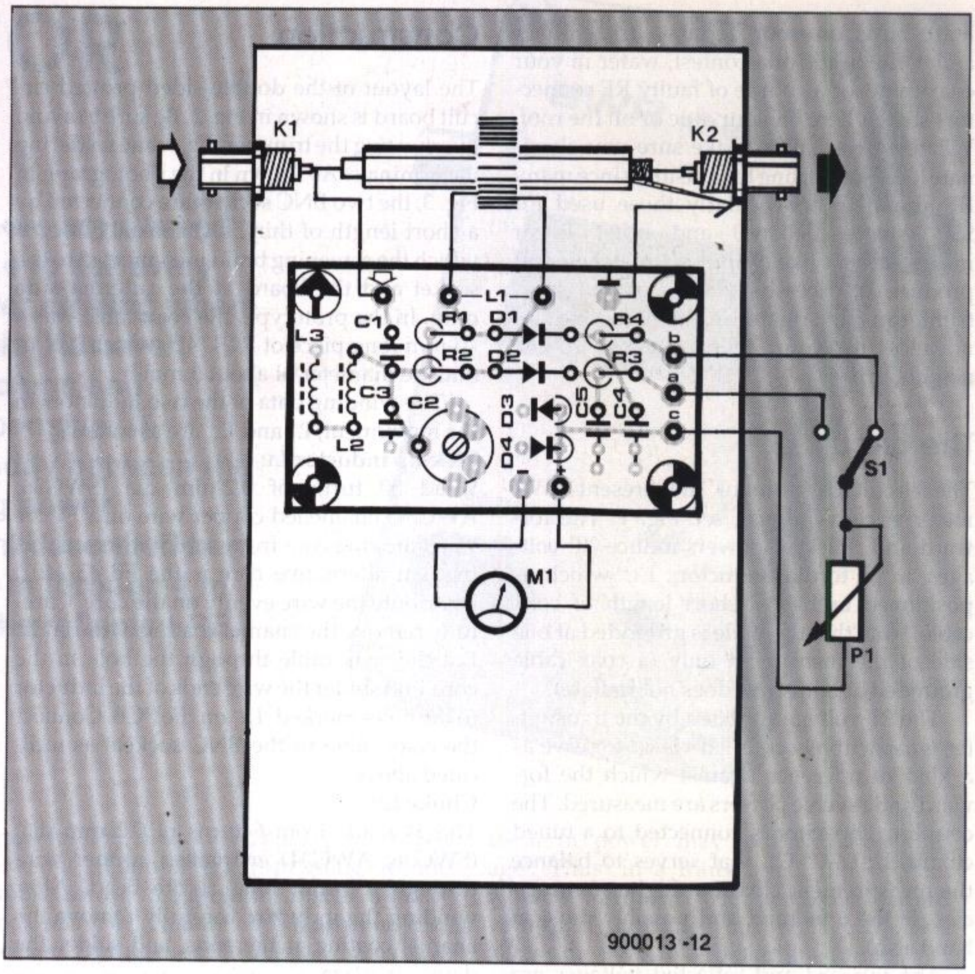
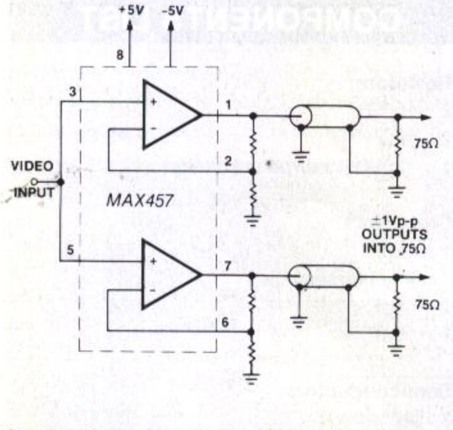


Fig. 4. Where to connect the external components.

NEW PRODUCTS

DUAL 70 MHz CMOS VIDEO AMPLIFIER



Maxim Integrated Products introduces the MAX457, a monolithic IC comprising two 70 MHz CMOS video amplifiers which can drive 75 Ω loads. These amplifiers operate from ±5 V supplies, and together consume only 350 mW. The only necessary external components are two resistors for gain setting, and two decoupling capacitors.

Without external compensation, these amplifiers are stable at a gain of 1 when driving 75 Ω loads, and stable at a gain of 2 when driving 150 Ω loads. This makes the MAX457 ideally suited to buffering video signals for transmission directly down 75 Ω cables, or down back-terminated 75 Ω cables. Other applications include driving flash converters, and providing output buffers for cross-point switches such as the MAX456.

Isolation between the two amplifiers is guaranteed to be greater than 60 dB. The MAX452 single version of this amplifier can be used in applications requiring even greater isolation.

Maxim Integrated Products • 120 San Gabriel Drive • Sunnyvale • CA 94086. Telephone: (408) 737-7600.

Maxim Integrated Products (UK) Ltd. • 21C Horseshoe Park • Pangbourne • Reading RG8 7JW. Telephone: (0734) 845255. Fax: (0734) 843863.

MORE SUPPORT SOFTWARE FOR ELEKTOR 8052-BASIC COMPUTER

RTB Computer Engineering now supply MCSCOM, a communication program specifically written to enable an IBM PC or compatible to support the Intel 8052-AH BASIC microcontroller. The main functions of MCSCOM are selected from a menu, and comprise:

A **dumb terminal** which communicates with the 8052 via a user selected port, using any of the standard PC baud rates. The X-on/X-off protocol is used. The modem control signals are ignored.

A **download** mode in which programs may be downloaded to the 8052. Only numbered lines are transmitted; un-numbered comments are ignored.

A **capture** mode which allows programs and other output from the 8052 to be captured and written to a disk file.

A **line editor** which is Wordstar compatible and may be used to modify program lines in the 8052. The line number itself may be edited to produce copies of a line if required.

Run-time parameters are stored in two files, one of which contains the user-defined defaults, the other the last used parameters. The parameter file may be selected from the command line at run time. This feature allows the user to carry on where he left off without having to redefine parameters.

For further details, prices and ordering information, send a self-addressed envelope to

R.T.B. Computer Engineering • 163 Duncombe Street • Sheffield S6 3RL. Telephone: (0742) 334370.

400-watt laboratory power supply

October 1989 and November 1990

A number of constructors of this popular project have brought the following problems to our attention.

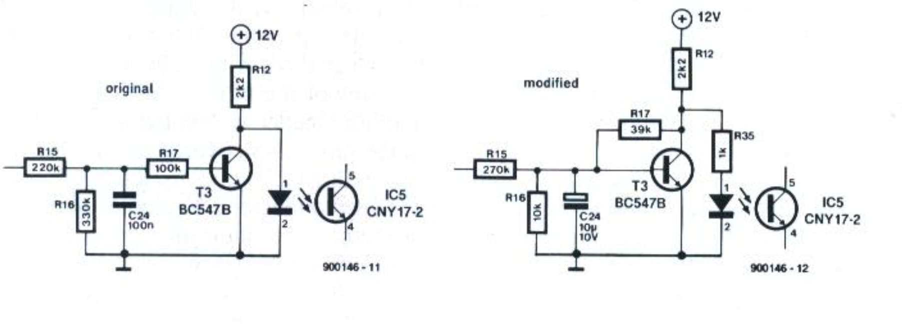
1. The onset point of the current limit circuit lies at about 3 A, which is too low. Solve this problem by replacing T1 with a Type BC517 darlington transistor, and R20 with a 82kΩ resistor.

2. Depending on the current transfer ratio of the optocoupler used, the transformer produces ticking noises. This effect, which is caused by overshoot in the pre-regulation circuit, may be traced with the aid of an oscilloscope monitoring the voltage across C26 at a moderate load current. The capacitor must be charged at each cycle of the mains frequency, and not once every five cycles. The problem is best solved by reducing the amplification of the regulation circuit. Replace R17 with a 39 kΩ resistor, and create feedback by fitting it between the base and the collector of T3. Also add a resistor in series with the optocoupler. These two changes are illustrated in Figs. 1 and 2. Lower R16 to 10 kΩ, increase C24 to 10 µF, and increase R15 to 270 kΩ.

3. Excessive heating of the transformer is caused by a d.c. component in the primary winding. This is simple to remedy by fitting a capacitor of any value between 47 nF and 470 nF, and a voltage rating of 630 V, across the primary connections. This capacitor is conveniently mounted on to the PCB terminal block that connects the transformer to the mains.

4. One final point: when using LED DVMs for the voltage/current indication, their ground line must be connected to the **positive** terminal of C12.

1



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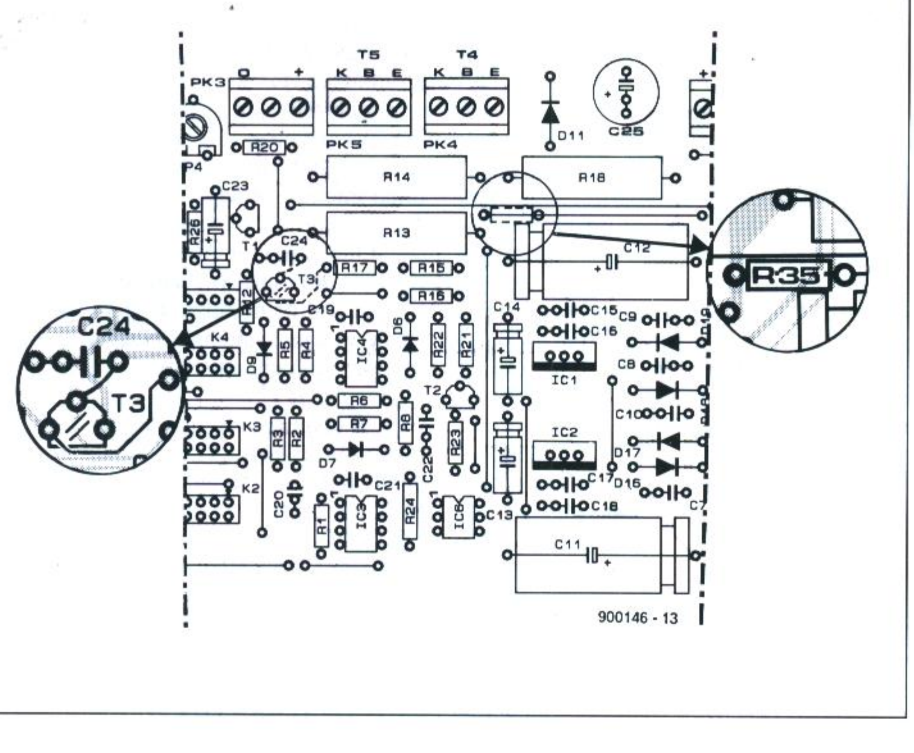

Hard disk monitor

December 1989

In some cases, the circuit will not reset properly because the CLEAR input of IC3A is erroneously connected to ground. Cut the ground track to pin 3 of IC3, and use a short wire to connect pin 3 to pin 16 (+5 V).

Microprocessor-controlled telephone exchange

October 1990

In some cases, the timing of the signals applied to IC17 causes a latch-up in the circuit, so that the exchange does not detect the state of the connected telephones properly. Solve this problem by cutting the track to pin 1 of IC17, and connecting pin 1 to ground (a suitable point is the lower terminal of C6).

The text on the fitting of wires on the BASIC computer board (page 19, towards the bottom of the right-hand column) should be modified to read: 'Finally, connect pin 6 of K2 to pin 7 of IC3 ($\overline{Y7}$ signal).'

S-VHS/CVBS-to-RGB converter (2)

October 1990

The capacitor marked 'C37', next to R21 on the component overlay (Fig. 7b and ready-made printed circuit board), should be marked 'C39'.

In case they are difficult to obtain locally, inductors type 119-LN-A3753 (L1) and 119-LN-A5783 (L2) may be replaced with the respective types 119-ANA-5874HM and 119-ANA-5871HM, also from Toko, Inc. Suggested suppliers are Cirkit Distribution Ltd., and C-I Electronics.

EPROM simulator

December 1989

Counters IC3 and IC4 may not function properly owing to a too low supply voltage. This problem may be solved by replacing IC12 with a 7806. Alternatively, use BAT85 diodes in positions D1 and D2.

Programmer for the 8751

November 1990

The ready-programmed 8751 for this project is available at £35.25 (plus VAT) under order number **ESS 7061**, not under order number ESS 5951 as stated on the Readers Services pages in the November and December 1990 issues.

Medium-power A.F. amplifier

October and November 1990

A circuit board to accommodate the smoothing capacitors in the symmetrical power supply section of the amplifier was designed after publication of Part 2 of this project. The track layout (mirror image) and the component mounting plan are given in Fig. 3 to assist readers in producing this PCB, which is not available ready-made. ■

3